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Application of Fuzzy Logic for Performance Enhancement of Drives

Gilberto Costa Sousa
University of Tennessee - Knoxville

To the Graduate Council:

I am submitting herewith a dissertation written by Gilberto Costa Sousa entitled "Application of Fuzzy Logic for Performance Enhancement of Drives." I have examined the final electronic copy of this dissertation for form and content and recommend that it be accepted in partial fulfillment of the requirements for the degree of Doctor of Philosophy, with a major in Engineering Science.

Bimal K. Bose, Major Professor

We have read this dissertation and recommend its acceptance:

Frederick W. Symonds, Walter L. Green, J. M. Bailey, B. R. Upadhyaya

Accepted for the Council:

Carolyn R. Hodges

Vice Provost and Dean of the Graduate School

(Original signatures are on file with official student records.)

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APPLICATION OF FUZZY LOGIC FOR PERFORMANCE ENHANCEMENT OF DRIVES

A Dissertation

Presented for the

Doctor of Philosophy

Degree

The University of Tennessee, Knoxville

Gilberto Costa D. Sousa

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ABSTRACT

Fuzzy logic shows enormous potential for advancing power electronics technology. Its application to DC and AC drives control is discussed here.

Initially, a phase-controlled bridge converter DC drive was considered. Analysis of converter performance at continuous and discontinuous conduction modes was first conducted. Fuzzy control was used to linearize the transfer characteristics of the converter in discontinuous conduction mode. It was then extended to current and speed loops, replacing the conventional proportional-integral controllers. The control algorithms were developed in detail, and verified by PC-SIMNON (developed by Lund Institute of Technology, Sweden) digital simulation. Significant performance improvement was achieved over conventional control methods.

Efficiency optimization of an indirect vector controlled induction motor drive was next considered. An accurate loss model of the converter induction machine system was first developed. Steady-state fundamental and harmonics loss characteristics, besides the dynamics of the machine were analyzed and incorporated in the model, resulting in a new synchronous frame dynamic D^e-Q^e equivalent circuit. The converter system has been modeled accurately for conduction and switching losses. The lossy models were then used in the validation of the fuzzy logic based on-line efficiency optimization control. At steady-state, the fuzzy controller adaptively changes the excitation current on the basis of measured input power, until the maximum efficiency point is reached. The pulsating

torque, due to flux reduction, has been compensated by an ingenious feedforward scheme. During transients, rated flux is established, to get the best transient response. After a comprehensive simulation study, an experimental 5 hp drive system was tested, with the proposed controller implemented on a Texas Instrument TMS320C25 digital signal processor, and the theoretical development was fully validated.

Finally, fuzzy logic was applied in combination with model-reference adaptive control (MRAC) technique to slip gain tuning of an indirect vector controlled induction motor drive. The MRAC methods based on reactive power and D-axis voltage were combined through a weighting factor, generated by a fuzzy controller, that ensures the use of the best method for any point in the torque-speed plane. A second fuzzy controller tunes the slip gain based on combined detuning error and its slope. The drive performance was extensively investigated through simulations and experiments. The results confirmed the validity of the proposed method.

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CHAPTER 1

INTRODUCTION

1.1 Artificial Intelligence and Adjustable Speed Drives

Adjustable speed drive technology has evolved enormously over the past 30 years. This evolution has been made possible because of technological advances in a number of related fields, such as power semiconductor devices, converter topologies and control techniques, microprocessors and digital signal processors (DSP). More recently, the application of intelligent control technologies, such expert systems, fuzzy logic and artificial neural networks, is advancing the frontier of this technology [19].

Among the artificial intelligence (AI) techniques, fuzzy logic is perhaps the most successful one, if judged from the standpoint of the number of practical applications [10]. Compared to other AI control techniques, i.e., expert systems and neural networks, fuzzy logic controllers numerically process structured knowledge that is embedded in only a few rules; expert systems symbolically process structured knowledge defined by a larger number of rules, whereas neural networks numerically process unstructured knowledge, and heavily depend on off-line learning techniques.

Fuzzy logic was introduced by Zadeh [1] in 1965, and can be viewed as a generalization of conventional Boolean logic. Its main purpose is the mathematical representation of imprecision and non-statistical uncertainty. The key concept in this

theory is that of a fuzzy set. A fuzzy set possesses a distinct feature of allowing partial membership, with degrees of membership varying from 0 (non-member) to 1 (full member). In contrast, in a "crisp" or Boolean set, a given element is a member of the set or it is not a member at all.

A fuzzy logic controller for a given process is capable of embedding in the control strategy, the qualitative knowledge and experience that an operator or engineer has about the process. Typically, the design of a fuzzy logic controller starts with a linguistic description of the control strategy, in the form of IF-THEN rules, that relate some process states (usually error and change in error), to the appropriate control action. The next step consists in making a quantitative interpretation of the linguistic variables, in the form of fuzzy sets. This is the most difficult step in the design process, and usually requires a significant number of iterations. The design process is completed by selecting an inference method, that essentially defines how the set of control rules, i.e., the rule base, is evaluated to derive the control action for a particular process state.

Fuzzy logic control does not require an accurate model of the plant, and therefore, it suits well to a process for which a formal model is unknown or ill-defined. It also works well for complex non-linear multi-dimensional systems, systems with parameter variation problem, or where the sensor signals are not precise.

Fuzzy Logic applications to power electronics and drives are almost entirely new, and this work is one of the first attempts to systematically apply fuzzy logic to the control of DC and AC drives. Prior to 1991, Li and Lau [5] applied fuzzy logic to a microprocessor-based servomotor controller, assuming a linear power amplifier. They also

compared system performance under PID control and model-reference adaptive control (MRAC) to that obtained with fuzzy control, and demonstrated the superiority of the latter. Da Silva et al [6] developed a fuzzy adaptive controller for a four quadrant power converter. The control system possesses a three-level hierarchical structure capable of identifying the best control law for any load condition, such that robust operation is achieved even under sudden changes in load.

Fuzzy logic shows enormous potential for advancing power electronics technology. Its ability to incorporate qualitative knowledge and to deal with imprecise information makes it very attractive to power electronics systems, where non-linearity is a common feature, and a precise model is difficult to obtain. In such cases, fuzzy logic can be applied to derive a linguistic model of the system, that describes its operational behavior. This fuzzy model can then be used in the design of a fuzzy controller for the system, just as a mathematical model is helpful in the design of a conventional controller.

1.2 Principles of Fuzzy Logic Control

It is appropriate to review here some basic concepts of fuzzy logic and fuzzy control. Although numerous fuzzy inference methods have been proposed in the literature [11]-[12], this discussion is based on Zadeh's [1]-[2] and Mamdani's [3] approaches.

A fuzzy set is characterized by a membership function, that associates each element in the set to a number in the real interval [0,1], that represents its grade of membership in the set. In contrast, a Boolean or "crisp" set is defined by a characteristic function, that maps all members to logic level 1 and the non-members to 0. Fig. 1.1

difference for the case of a hypothetical temperature control system. In Fig. 1.1(a), a crisp classification is provided, such that, the temperature value T = 67 °F is a member of the HOT set only. In contrast, in Fig. 1.1(b) temperature is considered a fuzzy variable, and T = 67 °F is a partial member of both MILD and HOT fuzzy sets. Because language is the primary vehicle for conveying knowledge, fuzzy variables are usually referred to as linguistic variables, and the fuzzy sets are viewed as the mathematical representation of their linguistic values (e.g., COLD, MILD, and HOT in Fig. 1.1). The numerical interval, that is relevant for the description of a fuzzy variable, is commonly named Universe of Discourse (in Fig. 1.1, the real interval [20,90]).

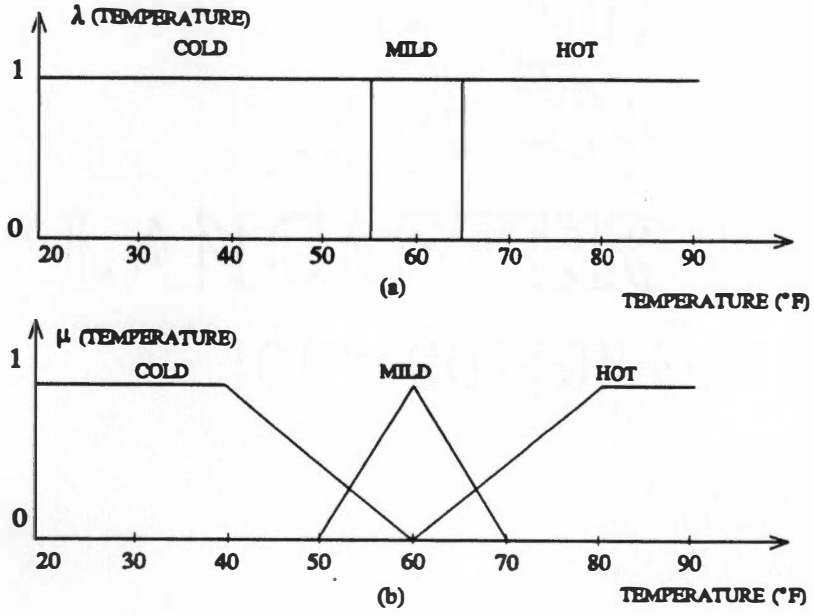


Fig. 1.1 Representation of temperature using:

(a) Crisp sets. (b) Fuzzy sets.

A few operations of Boolean set theory are also valid in fuzzy set theory, and are described below. Let $\mu_A(x)$ denote the degree of membership of a given element x in the fuzzy set A.

Union: Given two fuzzy sets A and B, defined on a universe of discourse X, the union ($A \cup B$) is also a fuzzy set of X, with membership function given as

$$\mu_{A \cup B}(x) = Max[\mu_A(x), \mu_B(x)] \qquad (1.1)$$

where x is any element of X.

Intersection: The intersection of two fuzzy sets A and B of the universe of discourse X, denoted by $A \cap B$, has the membership function given by

$$\mu_{A \cap B}(x) = Min[\mu_A(x), \mu_B(x)] \qquad (1.2)$$

Compliment: The compliment of a given set A of the universe of discourse X is denoted by $\neg A$, and has the membership function

$$\mu_{\neg A} = 1 - \mu_A(x) \qquad (1.3)$$

Fig. 1.2 illustrates these three basic properties. For fuzzy logic control, a few more concepts, such as fuzzy implication (fuzzy rules) and fuzzy composition (fuzzy inference) are important. A fuzzy rule typically has an IF-THEN format as follows:

IF (x is A AND y is B) THEN (z is C)

where x, y and z are fuzzy variables and A, B and C are fuzzy sets in the universes of discourse X, Y and Z, respectively. If the conditions expressed in the antecedent (IF portion) are met, then the action(s) specified in the consequent (THEN portion) are

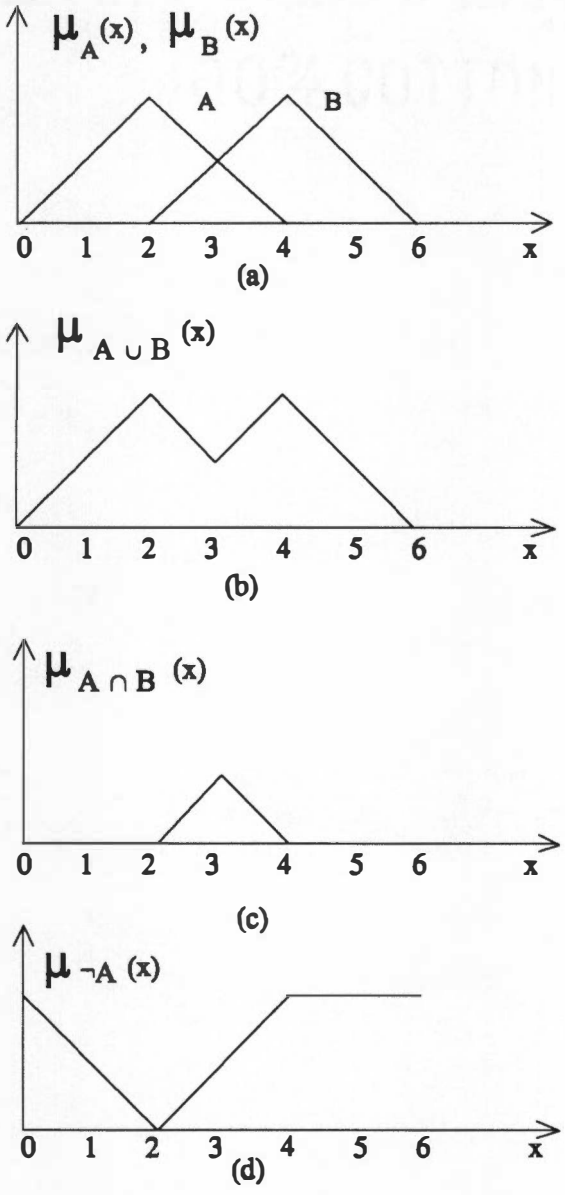


Fig. 1.2 Basic operations involving fuzzy sets.

(a) Original fuzzy sets defined on X.

(b) Union. (c) Intersection. (d) Negation.

taken. In order to design a fuzzy controller, a fuzzy rule base consisting of several rules must be constructed. For example, consider a hypothetical fuzzy speed control system for a DC motor, where the speed error (E) and change in error (CE) are used to determine changes in the control signal (DU), that in this case is the command armature current I_a^*. A part of the rule base would be:

Rule 1: IF E is Zero AND CE is Zero THEN DU is Zero;

Rule 2: IF E is Zero AND CE is Negative Small THEN DU is Negative Small;

Rule 3: IF E is Positive Small AND CE is Negative Small THEN DU is Zero;

Here, error (E), change in error (CE) and change in control (DU) are considered fuzzy variables, with possible values given by fuzzy sets such as Positive Small, Negative Small, and so on. As illustrated in Fig. 1.1, a given numerical value can be a member of more than one fuzzy set. This means that, for a particular input pair of values (E and CE), more than one rule could be activated or "fired". Therefore, there must be a way to combine the individual control actions of the fired rule, such that a single, meaningful action is taken. In fuzzy logic terms, the composition operation is the mechanism by which such task can be performed. Although several composition principles have been proposed in the literature, the most common one is the SUP-MIN (SUPremum-MINimum) composition. In a simplistic way, given a rule base, it is possible to construct a n-dimensional fuzzy relation R (consider it as a function of n variables). The simplest case is a single input (x), single output (u) system, resulting in a two-dimensional fuzzy relation, represented by a membership function $\mu_R(x,u)$. For this case, the composition operation can be expressed as:

$$\mu_B(u) = Sup_x\,[Min(\mu_A(x), \mu_R(x,u))] \qquad (1.4)$$

where A is the known fuzzy set for the input x and B is the inferred fuzzy set for the output u. In practice, the fuzzy relation R is seldom evaluated explicitly; instead, the SUP-MIN composition is applied to one rule at a time, and the individual control actions combined using the union operation. Fig. 1.3 illustrates the fuzzy composition by SUP-MIN principle for the two stated rules. Note that the output membership function of each rule is given by MIN operator whereas the combined fuzzy output is given by the SUP operator. This will be evident by the numerical example in Chapter 2.

The general structure of a fuzzy control system is given in Fig. 1.4. The control signal U is inferred from the two state variables error (e) and change in error (ce) (de/dt for the sampling interval). The e and ce are per unit (pu) signals derived from the actual E and CE signals by dividing with the respective gain factors as shown. In a strict sense, the fuzzy controller is designed to process fuzzy quantities only. Therefore, all crisp input values must be converted to fuzzy sets before being used. This process is called fuzzification operation, and can be performed by considering the crisp input values as "singletons" (fuzzy sets that have membership value of 1 for the given input value, and 0 for all other points in the universe of discourse). In Fig. 1.3, the given input values E' and CE' were "converted" to fuzzy sets by this process, before being compared to the other fuzzy sets. In a similar way, there is a need for converting the output of the fuzzy controller (a fuzzy set) to a crisp value required by the plant. This is called defuzzification operation, and can be performed by a number of methods of which the center-of-gravity (also known as centroid) and height methods are common. The centroid defuzzification

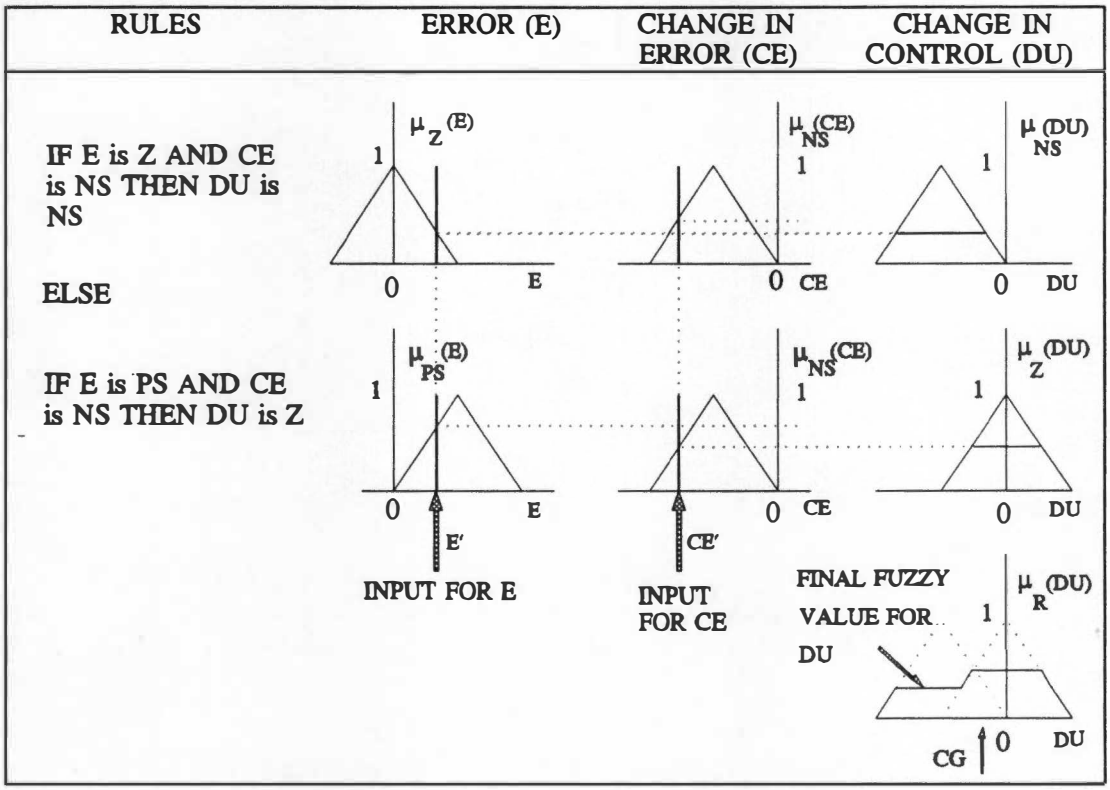


Fig. 1.3 Fuzzy composition method by SUP-MIN principle.

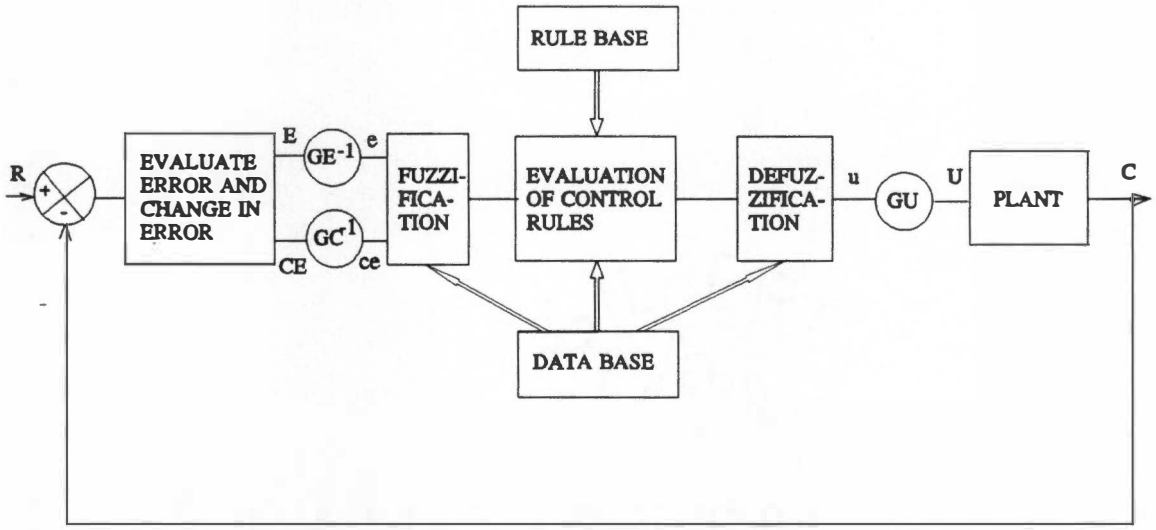


Fig. 1.4 Basic structure of a fuzzy controlled system.

method selects the output crisp value corresponding to the center of gravity of the output membership function, which is given by the expression

$$U_o = \frac{\int u\, \mu(u)\, du}{\int \mu(u)\, du} \tag{1.5}$$

In the height method, the centroid of each output membership function associated with every rule is first evaluated. The final output is then calculated as the average of the individual centroids, weighted by their heights (degree of membership) as

$$U_o = \frac{\sum_{i=1}^{n} u_i\, \mu(u_i)}{\sum_{i=1}^{n} \mu(u_i)} \tag{1.6}$$

Finally, the data base provides the operational definitions of the fuzzy sets used in the control rules, fuzzification and defuzzification operations.

Fuzzy logic controllers have been successfully applied to a number of different processes, in many cases yielding improved performance, compared to conventional control techniques. A qualitative explanation for this superior performance will now be discussed.

For a discrete PI controller, the change in control $\Delta u(k)$ can be expressed as

$$\Delta u(k) = K_p\, ce(k) + K_i\, \Delta t\, e(k) \tag{1.7}$$

where K_p and K_i are the proportional and integral gains, respectively, e(k) is the control error, ce(k) (=e(k) - e(k-1)) is the change in error, and Δt is the sampling interval. For a specific value of error e(k), the control increment $\Delta u(k)$ varies linearly with ce(k). In other

words, the mapping produced by a PI controller results in a plane in the three-dimensional space (ce, e, Δu). On the other hand, for a fuzzy controller, any type of mapping is theoretically possible. Again, for a given value of error e(k), the relationship between Δu(K) and ce(k) is not confined to a straight line, resulting in practice, in a variable K_p gain. Of course, in order to realize the potential benefits of this highly flexible structure, a much more elaborated design process is required. In response to this challenge, significant research effort has been devoted to the development of a theory for fuzzy dynamic systems, addressing problems such as fuzzy state variables, controllability and stability [12].

1.3 Outline of the Dissertation

The application of fuzzy logic to the control of both DC and AC drive systems is discussed in this work. Phase-controlled converter DC drives have been used in applications that usually require fast response, such as steel mills. Under normal conditions, the converter operates in a continuous conduction mode, in which the converter transfer characteristics are linear, and fast transient response for the current loop is obtained. However, at light load conditions, the converter tends to operate in a discontinuous conduction mode, that results in non-linear converter transfer characteristics and sluggish response of the current loop. Ultimately, the overall system performance is adversely affected. Furthermore, in many applications load disturbance is quite common, as well as parameter variation in the drive system. Consequently, fuzzy logic control can be applied with potential advantages over classical control techniques.

In Chapter 2, a speed control system that uses a phase-controlled bridge converter and a separately excited DC machine was investigated. Fuzzy control was used to linearize the transfer characteristics of the converter in discontinuous conduction mode occurring at high speed and light loads. The fuzzy control was then extended to the current and speed control loops, replacing the conventional proportional-integral (PI) control method. The compensation and control algorithms were developed in detail and verified by digital simulation using PC-SIMNON.

Vector control drives are becoming the industry standard for high performance applications [19]. Under vector control, an AC machine drive dynamic imitates that of a separately excited DC machine drive, with the advantages of ruggedness and higher torque to inertia ratio.

Most drive systems operate at rated flux, to ensure fast transient responses. However, at light loads, rated flux operation results in unnecessarily high core losses, thus impairing the efficiency of the drive. To address this problem, fuzzy logic was applied to the efficiency optimization of a speed control system, using an indirect vector controlled induction motor drive. In order to validate the new control scheme, as well as to predict system performance, an accurate loss model of the converter induction machine system was first developed, as discussed in Chapter 3. The model properly represents the loss phenomena as well as dynamic behavior of both machine and converter. After validation of the lossy models, they were used by the fuzzy logic based on-line efficiency optimization control, presented in Chapter 4. During efficiency optimization, the rotor flux is typically reduced, causing undesirable low frequency torque pulsations. To address this

problem, an innovative feedforward torque compensator is also described in Chapter 4. The drive system with the proposed efficiency optimization controller was simulated with the lossy models of converter and machine, and its performance was thoroughly investigated. An experimental 5 hp drive system, with the proposed controller implemented on a Texas Instruments TMS320C25 digital signal processor, was tested in the laboratory to validate the theoretical development.

Indirect vector control induction motor drives require the knowledge of rotor time constant, to generate the slip frequency used in the synthesis of the unit vectors. Fast dynamic response is only achieved when the slip gain used in the control matches the actual machine rotor time constant. Unfortunately, machine parameters are temperature and saturation dependent, and drive performance tends to deteriorate. Chapter 5 discusses the application of fuzzy logic to the slip gain tuning of an indirect vector controlled induction motor drive, using the standard model-reference adaptive control (MRAC) technique. The MRAC methods based on reactive power and synchronous frame D-axis voltage are combined together with a weighting factor that is generated by a fuzzy controller. The weighting factor ensures the dominant use of reactive power method in low speed high torque regions whereas the D - axis voltage method is dominant in high speed low torque regions. A second fuzzy controller tunes the slip gain based on the combined detuning error and its slope, so as to ensure fast convergence at any operating point in the torque-speed plane. Drive performance was extensively investigated, initially through PC-SIMNON simulation, and then using an experimental drive system.

CHAPTER 2

FUZZY CONTROL OF A DC DRIVE SYSTEM

2.1 Introduction

Direct current (dc) drives have been the traditional industry option in applications that require adjustable speed, mainly due to the ease of control and fast speed response. Although in the recent years alternate current (ac) drives have gained increased popularity, and will certainly dominate the adjustable speed drive market, dc drives still constitute the majority of industrial drive systems.

The application of fuzzy logic to the control of a dc servomotor has been discussed in the literature [5]. The authors described the system performance in the presence of sudden changes in inertia, under fuzzy control as well as under PID and MRAC control, and showed the superiority of fuzzy control. In their study, however, a linear power amplifier was assumed, and consequently, the problems associated with the non-linearities of actual power converters were not addressed.

Among the dc drive systems, the phase-controlled bridge converter drive shown in Fig. 2.1 is the most popular. It is shown here in a four-quadrant speed control mode, with inner current control loop, that provides fast transient response and also limits the armature current. Traditionally, proportional-integral (PI) control is used in both speed and current loops. The firing angle α is obtained from the current PI controller output V_s

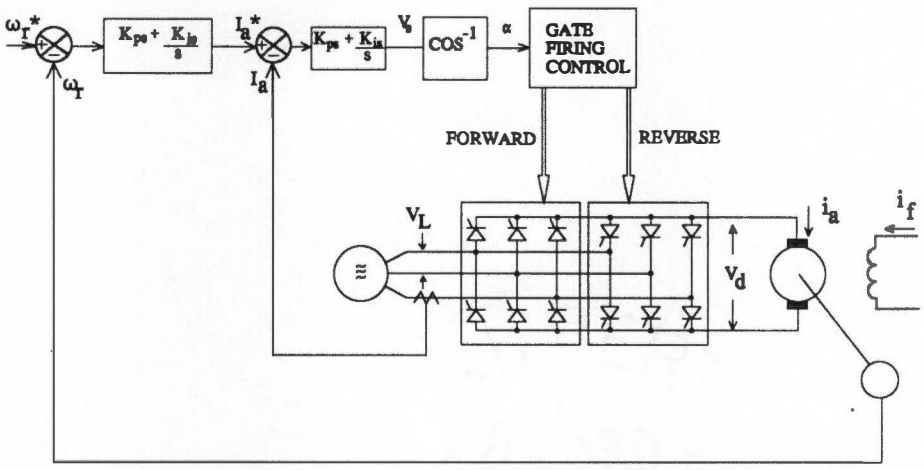


Fig. 2.1 Four quadrant phase-controlled converter DC drive.

by cosine wave crossing method, such that the converter output voltage V_d is proportional to V_s, i.e., a linear transfer characteristic is obtained for the converter system. The converter, however, may operate in either continuous or discontinuous conduction modes, as indicated by the current and voltage waveforms of Fig. 2.2. At high load torque, the difference between applied voltage V_d and the counter emf V_c is large enough such that the conduction will be continuous. However, at light load torques, this difference becomes small and the conduction tends to be discontinuous. Under discontinuous conduction, the converter characteristics is no longer linear, and the current loop response tends to deteriorate. This in turn results in a sluggish speed response, rendering the system inadequate for high performance applications. Among the number of methods suggested to linearize the converter transfer characteristics at discontinuous conduction mode, the look-up table method suggested by Ohmae et al. [15] appears to be very attractive. In this method, an auxiliary compensating $\Delta\alpha$ angle is generated as a function of main α angle and armature current I_a, and then added with α to generate the actual firing angle. The two-dimensional relation of $\Delta\alpha$ can be pre-computed for each X/R parameter and stored in the form of a look-up table for microcomputer implementation. If the parameter X/R variation is considered, and the compensating angle is needed with good accuracy, then the look-up tables memory tends to be very large.

The following sections discuss the application of fuzzy logic to a DC drive system. Initially, a fuzzy controller was designed to generated a $\Delta\alpha$ compensating angle, that when added to the original firing angle α, produces a linear characteristic for the converter system, even under discontinuous conduction mode. Next, fuzzy control was

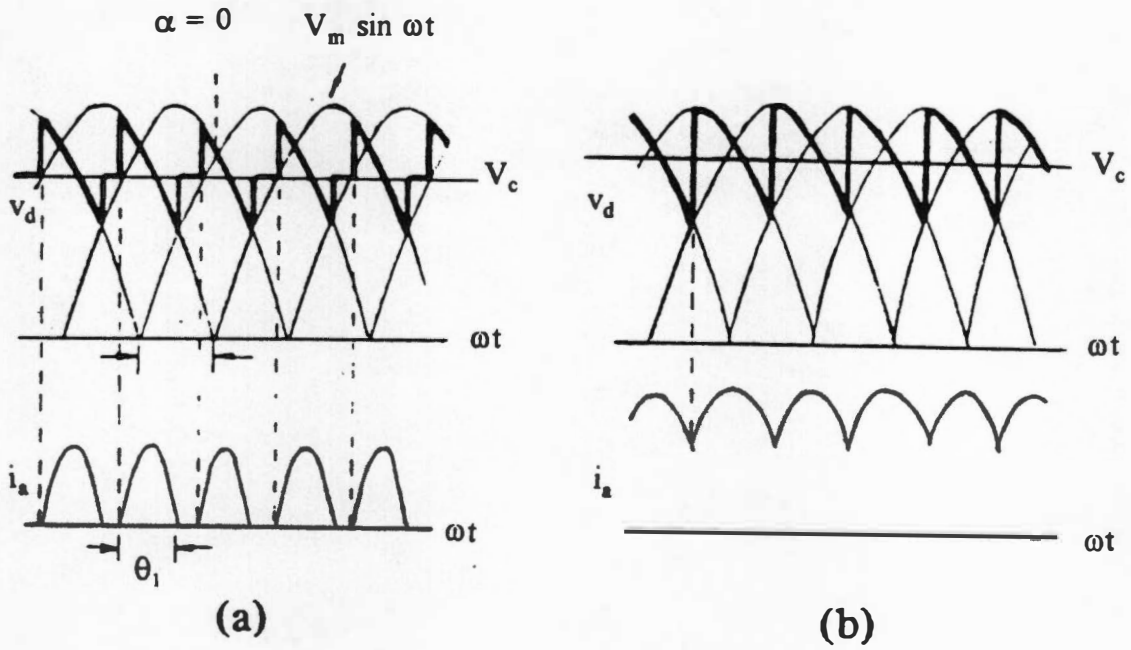


Fig. 2.2 Converter voltage and current waveforms [15].

(a) Discontinuous conduction.

(b) Continuous conduction.

extended to speed and current control loops, replacing the conventional PI control method. The compensation and control algorithms were developed in detail and verified by digital simulation of a drive system. The results were originally published at the IEEE IAS Annual Meeting, 1991 [14].

2.2 Fuzzy Controlled DC Drive System

The speed control system under consideration is shown in Fig. 2.3. The power circuit consists of a phase-controlled bridge converter that drives a separately excited DC motor. For simplicity, the converter is used in motoring mode only with fixed field excitation. Fuzzy controllers were used in both the speed control loop and in the inner current control loop. The current loop output V_s' is added with the estimated feed-forward counter emf signal $\hat{V}_c$, to generate the control signal V_s, which then generates the firing angle α, by cosine wave crossing method. The feed-forward addition of counter emf gives faster loop response. The fuzzy compensator is also indicated in the figure.

As mentioned in the introduction, the converter may operate in either continuous or discontinuous conduction mode. Complete mathematical analysis of converter performance characteristics has been made for both operating modes [21], and the relevant results will be repeated here. When current is flowing into the armature, the loop equation is given by

$$L\frac{di_a}{dt} + Ri_a = V_m \sin \omega t - V_c \qquad (2.1)$$

where V_c = machine counter emf, V_m = peak supply line voltage, i_a =

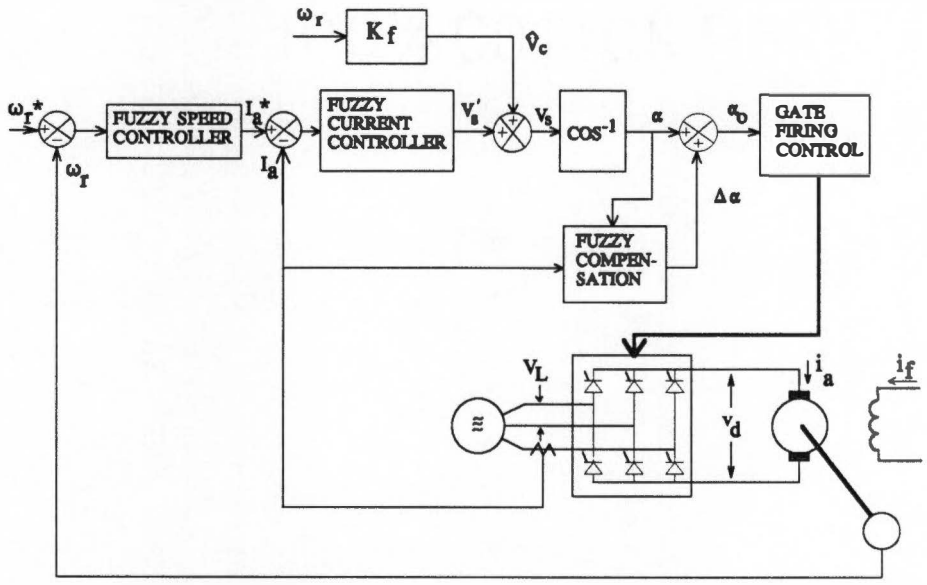


Fig. 2.3 Fuzzy controlled DC drive system.

instantaneous armature current, and R and L are , respectively, the resistance and inductance of the armature plus ac source and connections. In continuous conduction mode, the average armature current I_a and converter output voltage V_d, in normalized form can be given as follows

$$I_a(pu) = \frac{I_a}{3V_m/\pi X} = \frac{X}{R}\left[\cos\alpha - \frac{\pi V_c}{3V_m}\right] \tag{2.2}$$

$$V_d(pu) = \frac{V_d}{V_m} = \frac{3}{\pi}\cos\alpha \tag{2.3}$$

where X is the armature reactance (ωL), and α is the converter firing angle. The peak line voltage (V_m) can essentially be considered a constant, and therefore, V_d can be controlled linearly by V_s with cosine wave crossing technique, as indicated in Fig. 2.3. In discontinuous conduction mode, the following armature circuit equations are valid

$$I_a(pu) = \frac{I_a}{3V_m/(\pi X)} = \frac{X}{R}\left[\cos(\frac{\pi}{3}+\alpha) - \cos(\frac{\pi}{3}+\alpha+\theta_1) - \frac{V_c}{V_m}\theta_1\right] \tag{2.4}$$

$$V_d(pu) = \frac{V_d}{V_m} = \frac{3}{\pi}\left[\cos(\frac{\pi}{3}+\alpha) - \cos(\frac{\pi}{3}+\alpha+\theta_1) - \frac{V_c}{V_m}\theta_1\right] + \frac{V_c}{V_m} \tag{2.5}$$

$$\frac{V_c}{V_m} = \frac{\sqrt{1+(X/R)^2}}{\left[1-\exp(-\frac{R}{X}\theta_1)\right]}\left\{\sin(\frac{\pi}{3}+\alpha+\theta_1-\phi)-\sin(\frac{\pi}{3}+\alpha-\phi)\exp(-\frac{R\theta_1}{X})\right\} \tag{2.6}$$

where θ_1 is the conduction angle of current pulse ($0 < \theta_1 < \pi/3$), as indicated in Fig.2.2(a), and $\phi = \tan^{-1}$ (X/R). For a fixed X/R parameter, the equations 2.2 - 2.6 were plotted in Fig. 2.4 for different α angles, which also indicates the boundary between continuous and discontinuous conduction modes. For example, at $\alpha = 80°$, the conduction is continuous at the point A. As the machine counter emf is increased, the V_d (pu) remains constant at decreasing I_a (pu) until point B, when the conduction becomes discontinuous. Further increase of counter emf will cause increase of V_d (pu) until it reaches the point at which I_a(pu) $\cong$ 0, if the machine is not loaded. The nonlinear V_d(pu) - I_a(pu) relation adversely affects the gain characteristics of the current control loop. If, for example, the loop gain is made optimum at continuous conduction mode, the lower gain at discontinuous conduction will make the loop response sluggish. On the other hand, if the gain is optimized for discontinuous mode at a certain operating point, the loop will tend to be unstable at continuous conduction. This problem can be overcome through Fuzzy $\Delta\alpha$ compensation. For instance, at operating point C in Fig. 2.4, if the proper $\Delta\alpha$ angle is added to α_o=80° , the compensated V_d moves to point D, that lies in the extension of the horizontal characteristics of the continuous conduction mode. As a consequence, the converter gain becomes the same as in the continuous conduction mode, and proper tuning of the current controller is now possible.

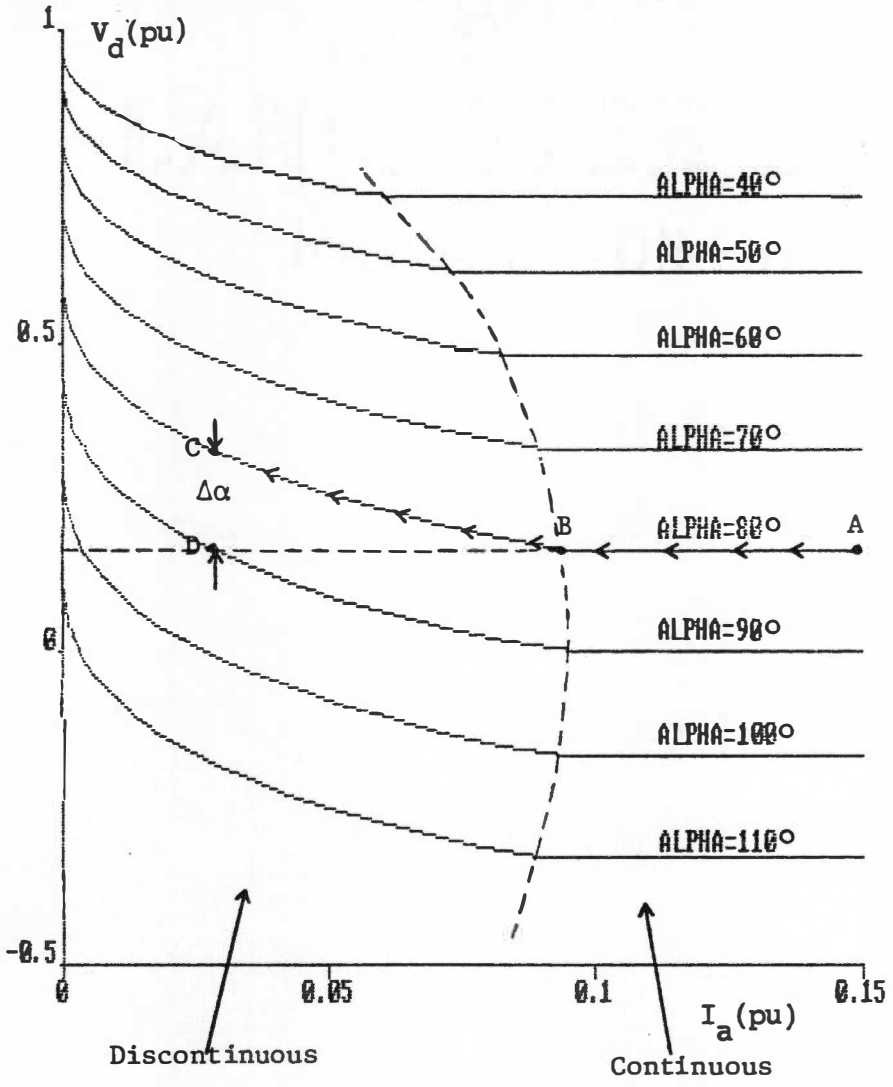


Fig. 2.4 Theoretical V_d-I_a (pu) phase plot without compensation.

2.2.1 Fuzzy Linearization of Converter Characteristics

It was mentioned in Chapter 1 that fuzzy control is well-suited in a non-linear system, especially where parameter variation problem exists. Therefore, we applied fuzzy method of $\Delta\alpha$ angle compensation, in order to linearize the converter transfer characteristics in discontinuous conduction mode. The special feature in fuzzy control is that the $\Delta\alpha$ angle is expressed as a fuzzy relation of the variables I_a and α angle. The set of rules for fuzzy compensation is given in matrix form in Table 2.1, where all the symbols are defined in the usual fuzzy logic terminology. A typical rule has the following structure:

IF I_a is small negative (NS) AND α is small positive (PS)

THEN $\Delta\alpha$ is small negative (NS)

The rule base was developed by heuristic from the viewpoint of practical system operation. The current I_a was represented in normalized form. Fig. 2.5 shows the membership function plots of the variables α, I_a(pu) and $\Delta\alpha$. The universes of discourse of the variables cover the whole discontinuous conduction region. The sensitivity of a variable determines the number of fuzzy sets required to describe it. The universe of discourse of α is described by five fuzzy sets, whereas I_a(pu) and $\Delta\alpha$ are described by nine and eleven sets, respectively. The linguistic terms used for the sets are for convenience only and must be interpreted in a "context free" grammar, since their conventional meaning does not correspond to the sign and numerical values of the variables. In Fig. 2.5, 50% overlap was provided for the neighboring fuzzy sets. Therefore, at any given point of the universe of discourse, no more than two fuzzy sets

Table 2.1 Rule base for $\Delta\alpha$ compensation.

I_a \ α	NB	NS	Z	PS	PB
NVB	NVB	PB	PB	PB	PB
NB	NVB	Z	Z	Z	Z
NM	NVB	NS	NVS	NVS	NVS
NS	NVB	NM	NS	NS	NS
Z	NVB	NB	NM	NM	NS
PS	NVB	NVB	NB	NM	NM
PM	NVB	NVB	NB	NB	NB
PB	NVB	NVB	NVB	NB	NB
PVB	NVB	NVB	NVB	NVB	NB

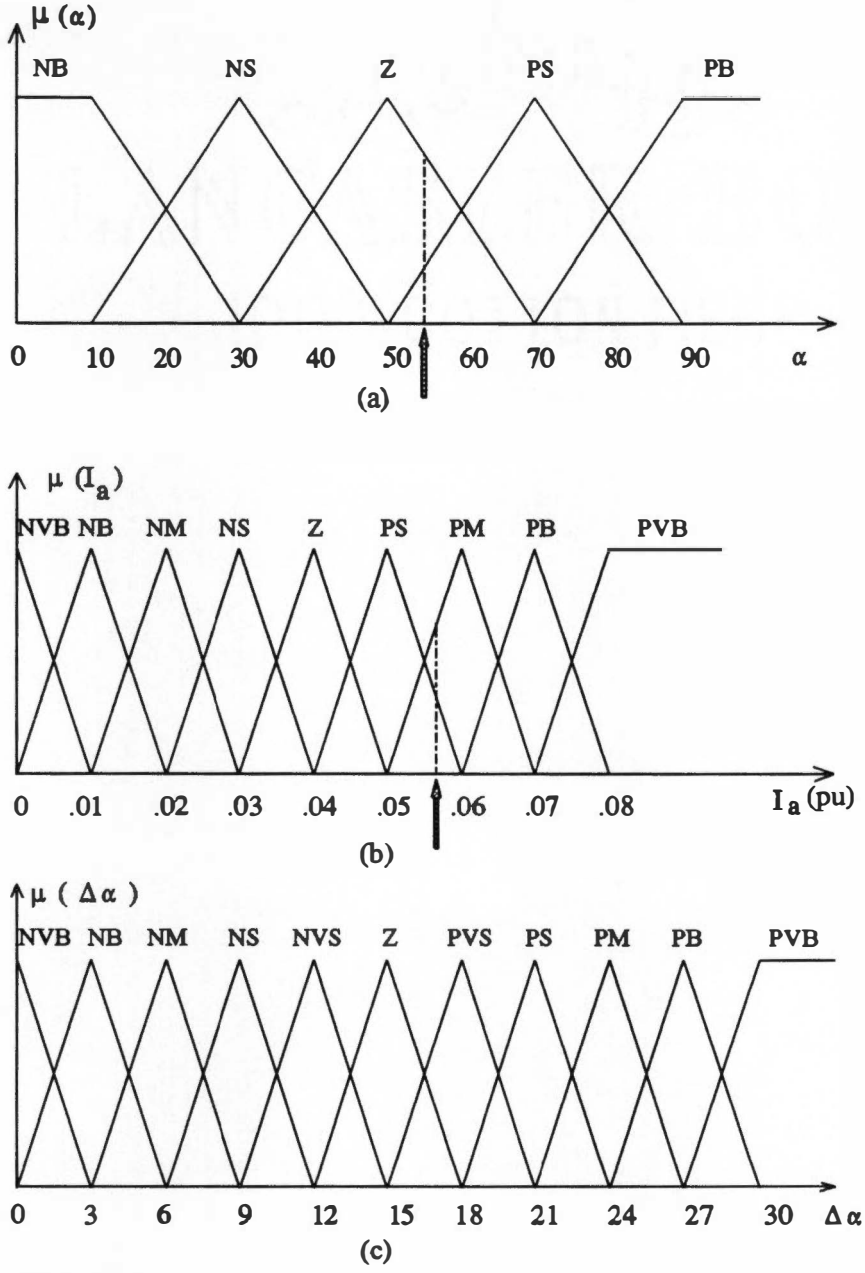


Fig. 2.5 Membership functions for $\Delta\alpha$ compensation.

a) Firing angle (α) b) Armature current (I_a)

c) Compensating angle ($\Delta\alpha$)

have non-zero degree of membership. This choice of fuzzy partitioning along with the SUP-MIN composition method resulted in a simplification of the fuzzy linearization algorithm. It is evident that, for any input data of I_a(pu) and α, at the most four rules will be valid in the entire rule base given in Table 2.1. The algorithm for fuzzy linearization is described below, along with a numerical example, included for clarity.

1. Sample the DC current I_a and firing angle α from the current control loop. Convert I_a to I_a(pu).

   [Lets assume: I_a = 5.6 A, α = 55°, I_{base} = 98.5 A (see Table 2.1)

   I_a(pu) = 5.6/98.5 = 0.057]

2. Calculate the interval indices I and J (that identify the interval number in the fuzzy sets) for α and I_a(pu), respectively as follows:

   I = INT ((α + 10)/20)

   J = INT ((I_a(pu) + 0.01)/0.01)

   [I = 3, J = 6]

3. Calculate the degree of membership of α and I_a(pu) for the left most fuzzy subset using I and J, respectively as follows:

   $\mu_Z(\alpha)$ = (20 I + 10 - α) / 20

   $\mu_{PS}(I_a(pu))$ = (0.01 I - I_a(pu))/0.01

   [$\mu_Z(55°)$ = 0.75, $\mu_{PS}(0.057)$ = 0.3]

4. Evaluate degree of membership for other sets by complement relation

   $\mu_{PS}(\alpha) = 1 - \mu_Z(\alpha)$

   $\mu_{PM}(Ia(pu)) = 1 - \mu_{PS}(I_a(pu))$

[$\mu_{PS}(55°) = 0.25$, $\mu_{PM}(0.057) = 0.7$]

5. Identify the four valid rules in Table 2.1 (stored as look-up table) and calculate the degree of membership μ_{Ri} contributed by each rule R_i [i = 1, 2, 3, 4], using MIN operator

   [$\mu_{R1} = \text{Min } \{\mu_Z(\alpha), \mu_{PS}(I_a(pu))\} = \text{Min } \{0.75, 0.3\} = 0.3$]

   [$\mu_{R2} = \text{Min } \{\mu_{PS}(\alpha), \mu_{PS}(I_a(pu))\} = \text{Min } \{0.25, 0.3\} = 0.25$]

   [$\mu_{R3} = \text{Min } \{\mu_Z(\alpha), \mu_{PM}(I_a(pu))\} = \text{Min } \{0.75, 0.7\} = 0.7$]

   [$\mu_{R4} = \text{Min } \{\mu_{PS}(\alpha), \mu_{PM}(I_a(pu))\} = \text{Min } \{0.25, 0.7\} = 0.25$]

6. Retrieve the amount of correction $\Delta\alpha_i$, i = 1,2,3,4 corresponding to each rule, from Table 2.1

   $\Delta\alpha_1 = (\alpha = Z, I_a(pu) = PS) \rightarrow \Delta\alpha_1 = NB = 3°$

   $\Delta\alpha_2 = (\alpha = PS\ I_a(pu) = PS) \rightarrow \Delta\alpha_2 = NM = 6°$

   $\Delta\alpha_3 = (\alpha = Z, I_a(pu) = PM) \rightarrow \Delta\alpha_3 = NB = 3°$

   $\Delta\alpha_4 = (\alpha = PS, I_a(pu) = PM) \rightarrow \Delta\alpha_4 = NB = 3°$

7. Calculate the crisp value of $\Delta\alpha$ by height defuzzification method as follows:

   [$\Delta\alpha = (0.30*3 + 0.25*6 + 0.70*3 + 0.25*3) / (0.30 + 0.25 + 0.70 + 0.25) = 3.5$]

The strength of fuzzy compensation is that the number of rules required to express the fuzzy relation is fairly small and memory requirement is low compared to large look-up table needed in conventional method.

2.2.2 Fuzzy Control of Current and Speed Loops

In addition to converter linearization, fuzzy logic control was applied to the speed and current loops as well, replacing the conventional PI controllers. The objective was to explore the control robustness in the presence of parameter variation and load disturbance effect. However, both loops must satisfy the requirements of fast transient response with minimum overshoot. With converter linearization, both speed and current loops have essentially first order characteristics. Therefore, intuitively the same fuzzy control strategy should be valid for both.

The selected fuzzy controller structure possesses two input variables, namely error (E) and change in error (CE), and one output variable, the change in control setting (ΔU). Such controller can be viewed as a generalization of the conventional PI controller, where the effective gains K_p and K_i are dependent on the input state (E,CE). The input variables considered in the fuzzy rule base are defined as:

$$E(k) = R(k) - C(k)$$

$$CE(k) = E(k) - E(k-1)$$

where R(k) = reference signal, C(k) = output signal and k = sampling interval. The control strategy is described linguistically by rules of the following format:

IF E(k) is X AND CE(k) is Y THEN $\Delta U(k)$ is Z

where $\Delta U(k)$ is the change in the control setting, X, Y and Z are the fuzzy sets defined in the universe of discourse of E, CE and ΔU, respectively. The variables can be expressed as per unit quantities as follows:

$$e(pu) = E(k)/GE$$

$$ce(pu) = CE(k)/GCE$$

$$\delta u(pu) = \Delta U(k)/GU$$

where GE, GCE and GU are the respective gain factors of the controller. The gain factors are normally different for speed and current control loops. The representation of the variables in terms of per unit values permits flexibility in the design and tuning of the controller. Fig. 2.6 shows the membership functions of e(pu), ce(pu) and δu(pu) variables. Note that the fuzzy sets for each variable have asymmetrical shape causing more crowding near the origin. This permits precision control near the steady state operating point, without unduly increasing the number of sets. However, a finer partitioning for δu(pu) was necessary considering the sensitivity of this variable. As fuzzy controller design is based on intuition and experience, instead of the system model, the following considerations were given in the beginning:

1. If both e(pu) and ce(pu) are zero, then maintain the present control setting U (k) (i.e., δu(pu) = 0);
2. If e(pu) is not zero, but is approaching this value at satisfactory rate, then maintain the present control setting U(k);
3. If e(pu) is growing, then change in the control signal δu(pu) is not zero and its value depends on the magnitude and sign of e(pu) and ce(pu) signals [7].

Table 2.2 gives the rule base matrix for current and speed controllers. A close look at the rule base indicates that the auxiliary diagonal consists of Z (zero) elements which conform to the second consideration as given above. Note that the value assigned to δu(pu) depends on the distance from the auxiliary diagonal. For instance, if e=PS and

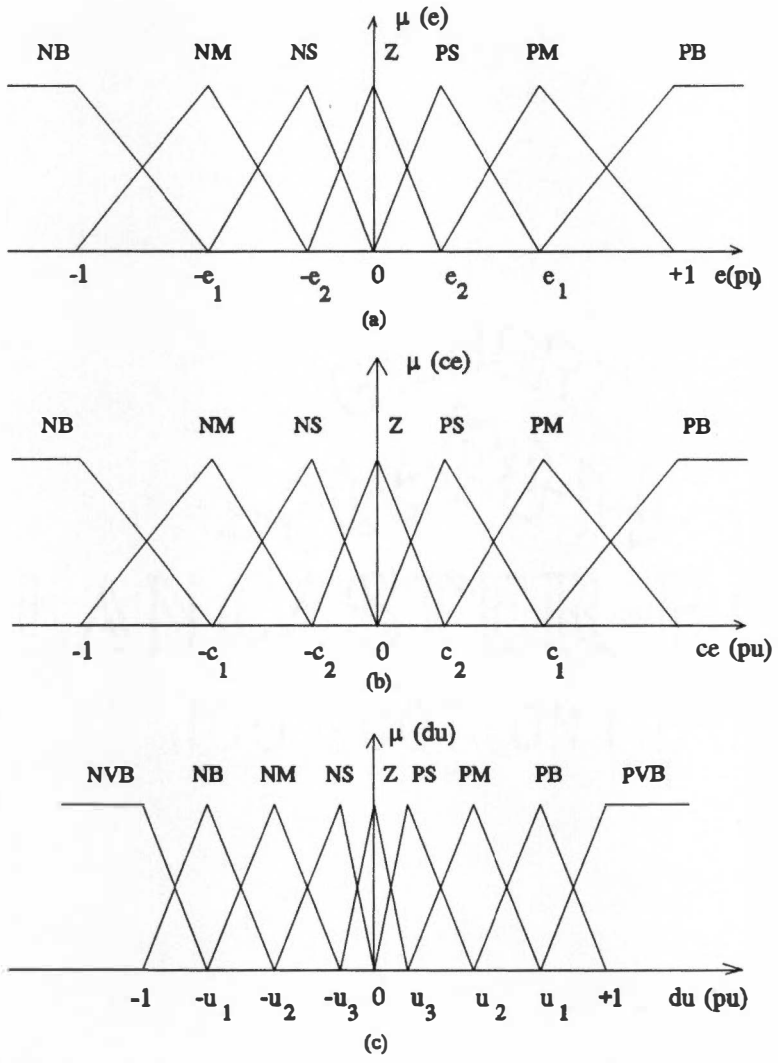


Fig. 2.6 Membership functions of speed and current controllers.

a) Error b) Change in error

c) Change in control

ce=NB, the system is approaching the steady state point (e=Z, ce=Z) too fast, and δu is made NM to prevent a large overshoot. On the other hand, if e=PS but ce=NS, the system is converging to the steady state point at the desired rate, and no change in control is required, i.e., δu=Z. The control strategy is dominantly affected by changes in the rule base, and these constitute the primary means for controller tuning.

Table 2.2 Rule base for current and speed controllers.

ce \ e	NB	NM	NS	Z	PS	PM	PB
NB	NVB	NVB	NB	NB	NM	NS	Z
NM	NB	NB	NM	NM	NS	Z	PS
NS	NB	NM	NS	NS	Z	PS	PM
Z	NB	NM	NS	Z	PS	PM	PB
PS	NM	NS	Z	PS	PS	PM	PB
PM	NS	Z	PS	PM	PM	PM	PB
PB	Z	PS	PM	PB	PM	PVB	PVB

A finer tuning was obtained by iterating the parameters e_1, e_2, ..., c_1, c_2, ..., u_1,u_2, ... in Fig. 2.6, until the desired controller performance was obtained. The control procedure is essentially identical to that of $\Delta\alpha$ compensation scheme. The steps for speed control can be summarized as follows:

1. Sample ω_r^* and ω_r.
2. Compute error (E), change in error (CE) and their per unit values as follows:

   $E(k) = \omega_r^*(k) - \omega_r(k)$

   $CE(k) = E(k) - E(k-1)$

   $e(pu) = E(k)/GE$

   $ce(pu) = CE(k)/GCE$
3. Identify the interval indices I and J (1, 2, etc. shown in Fig. 2.6) for e(pu) and ce(pu), respectively, by comparison method.
4. Compute the degree of membership of e(pu) and ce(pu) for the relevant fuzzy sets.
5. Identify the four valid rules in Table 2.2 and calculate the degree of membership μ_{Ri} using MIN operator.
6. Retrieve δu_i for each rule from Table 2.2
7. Calculate the resultant crispy value of $\delta u(pu)$ by height defuzzification method.
8. Compute the next control signal as

   $U(k) = U(k-1) + GU * \delta u(pu)$

The control for the current loop is the same as above, except here the gain factors GE, GCE and GU are different.

2.3 Simulation Study

In order to validate the control strategies discussed above, digital simulation studies were made using PC-SIMNON language. The phase-controlled rectifier and DC machine SIMNON models were initially developed and tested. Next, the fuzzy $\Delta\alpha$ compensator was constructed, along with the current and speed fuzzy controllers, and integrate to the other sub-systems. All SIMNON routines are listed in the appendix. Although SIMNON is well suited for linear and non-linear systems simulation, it does not support branch instructions, what complicates the programming of fuzzy controllers. A complete description of the simulation methodology will be postponed to Chapter 4.

Table 2.3 shows the parameters of the drive system used in the simulation study. The speed and current loops of the drive were also designed and simulated with PI control, in order to compare performance with the respective fuzzy control loop. The compensation and feedback control algorithms were iterated until best simulation results were obtained.

In the beginning, the performance of the fuzzy compensation scheme was tested keeping both the speed and current loops open. Fig. 2.7 shows the voltage and current responses without compensation. Initially, V_s was set such that $\alpha = 70°$, and then the drive simulation was enabled. With inertia load, the machine speed builds up freely with the developed torque. Initially, the conduction is continuous and $V_d = V_d^* = KV_s$. As speed builds up, the higher counter emf forces the converter to enter into discontinuous conduction, and V_d rises above the reference value V_d^* as shown. The experiment was then repeated with fuzzy $\Delta\alpha$ compensation enabled, for different values of α angle.

Table 2.3 Parameters of DC drive system.

DC Motor:		
110 V	2.5 hp	1800 rpm
R_a= 0.6 Ω	L_a= 8 mH	
J= 0.0465 Kg.m^2	B= 0.004 N.m.sec/rad	
K_t= 0.55 V.sec/rad		
I_a= 20 A (rated value)		
V_L= 90 V (line voltage from transformer)		
$T_L = K_L.\omega_r^2$ (load torque)		
K_L= 2.78 10^{-4} N.m.sec^2/rad		

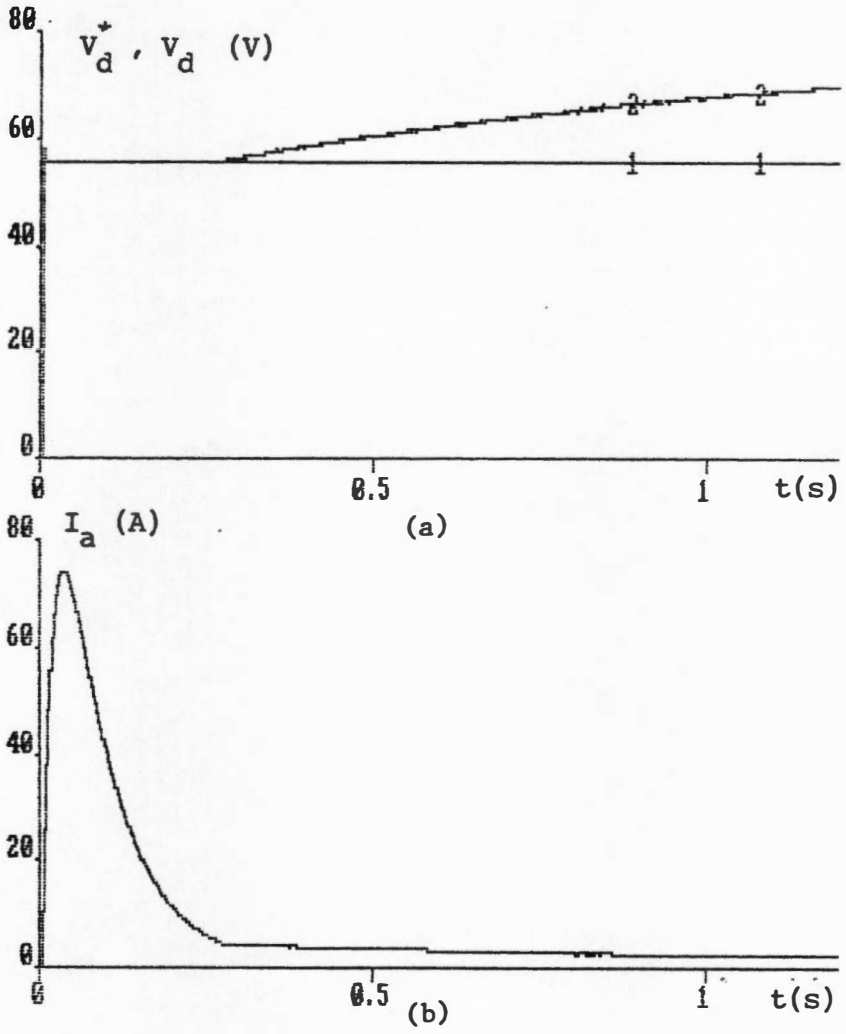


Fig. 2.7 Open loop V_d step (α = 70°) response without compensation.

a) Reference (1) and actual (2) dc voltages.

b) Armature current (dc value).

The corresponding V_d(pu) - I_a(pu) phase plots are given in Fig. 2.8. It is evident that with fuzzy $\Delta\alpha$ compensation, V_d(pu) at a certain angle essentially remains constant for the whole region of I_a(pu), i.e., V_d is essentially equal to V_d^* in both continuous and discontinuous regions. Fig. 2.9 shows the corresponding time domain plot of converter DC voltage (v_d) and armature current (i_a) waves with $\Delta\alpha$ compensation for $\alpha = 70°$, where time starts at 0.46 sec.

The current control loop was then tested with the fuzzy controller with the speed loop remaining open, but the speed was locked to a fixed value by assuming a very large inertia, so as to establish the discontinuous conduction mode. Fig. 2.10(a) shows the current loop response with $\Delta\alpha$ compensation which can be compared with Fig. 2.10(b) that gives response without compensation. In both cases, actual current waveform is shown. The boost of transient response due to converter linearization is evident. The responses of the current loop with PI control, with and without $\Delta\alpha$ compensation, were obtained and shown in Fig. 2.11 for comparison purpose. The response improvement in Fig. 2.10 for either condition indicates the superiority of fuzzy control.

Next, the speed loop was closed, and transient response was tested with both fuzzy current and speed control at linearized converter condition. Fig. 2.12 shows the speed and current response that covers both continuous and discontinuous regions. The figure also shows the effect of 40% step load torque applied at 0.8 sec. Fig. 2.13 shows the

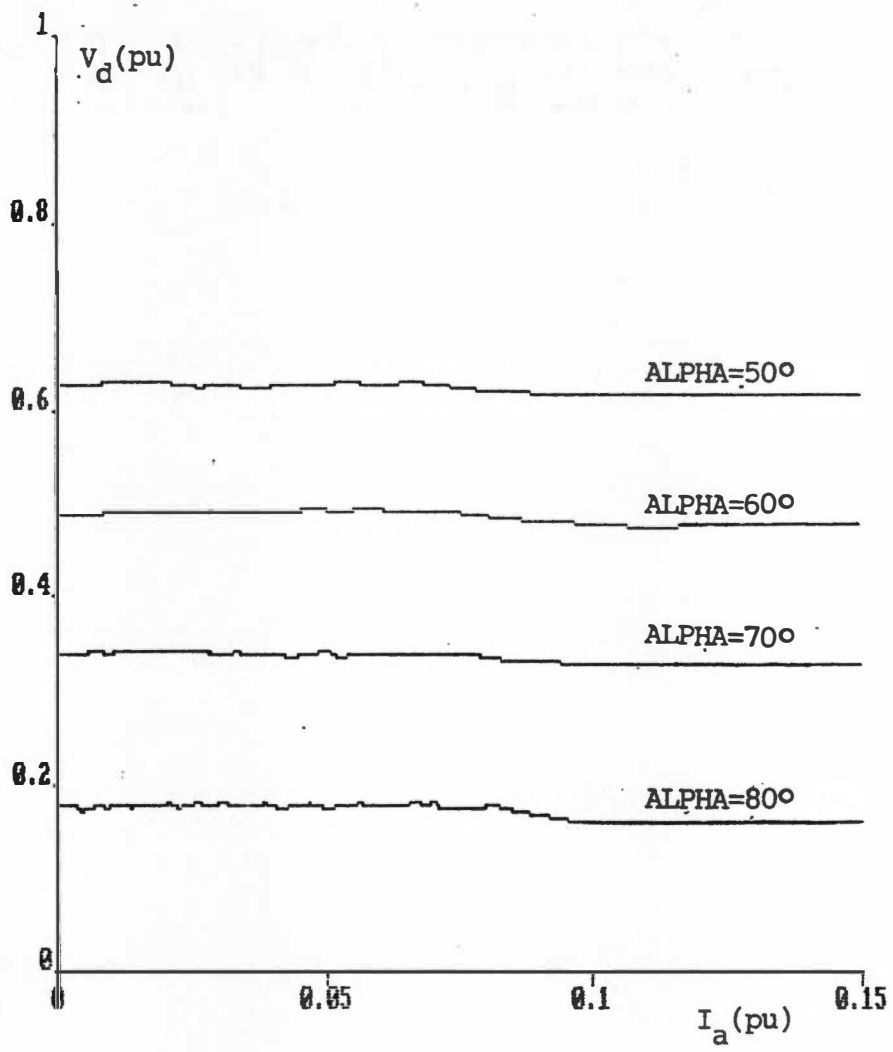


Fig. 2.8 Phase plot V_d-I_a with $\Delta\alpha$ compensation.

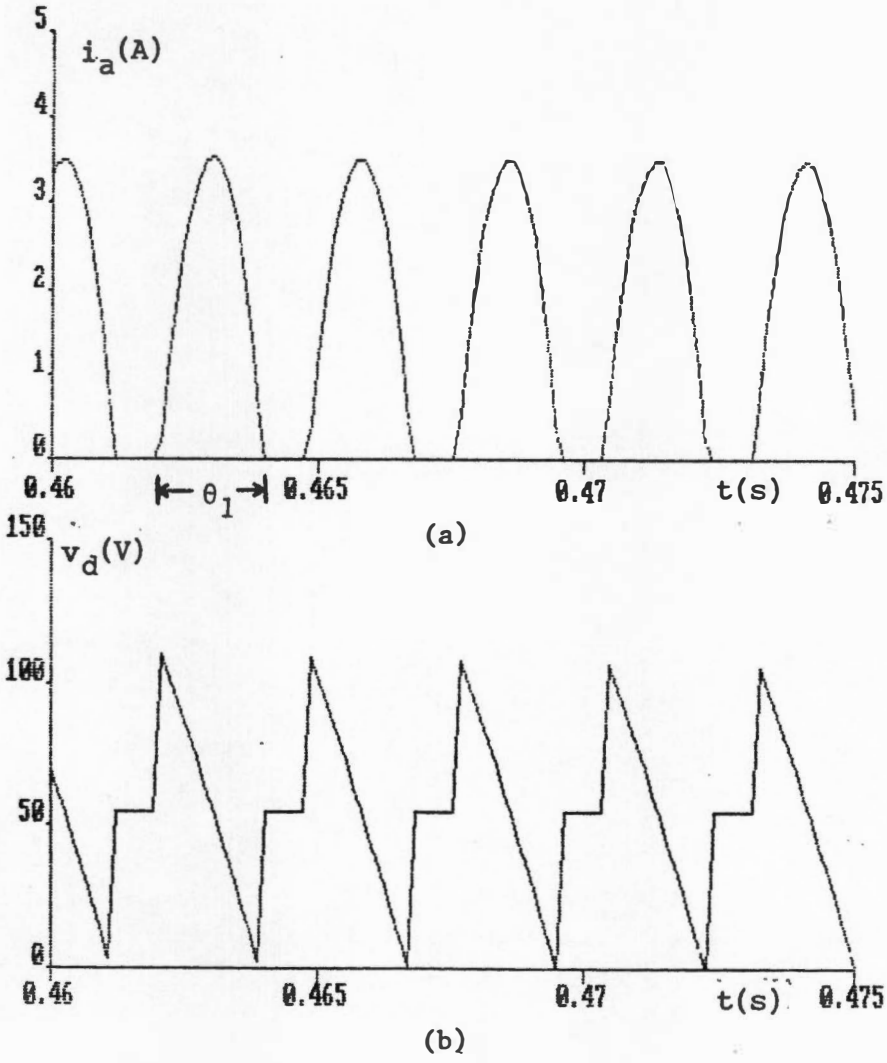


Fig. 2.9 Open loop V_d step response with $\Delta\alpha$ compensation.

a) Armature current (actual value).

b) Converter output voltage (actual value).

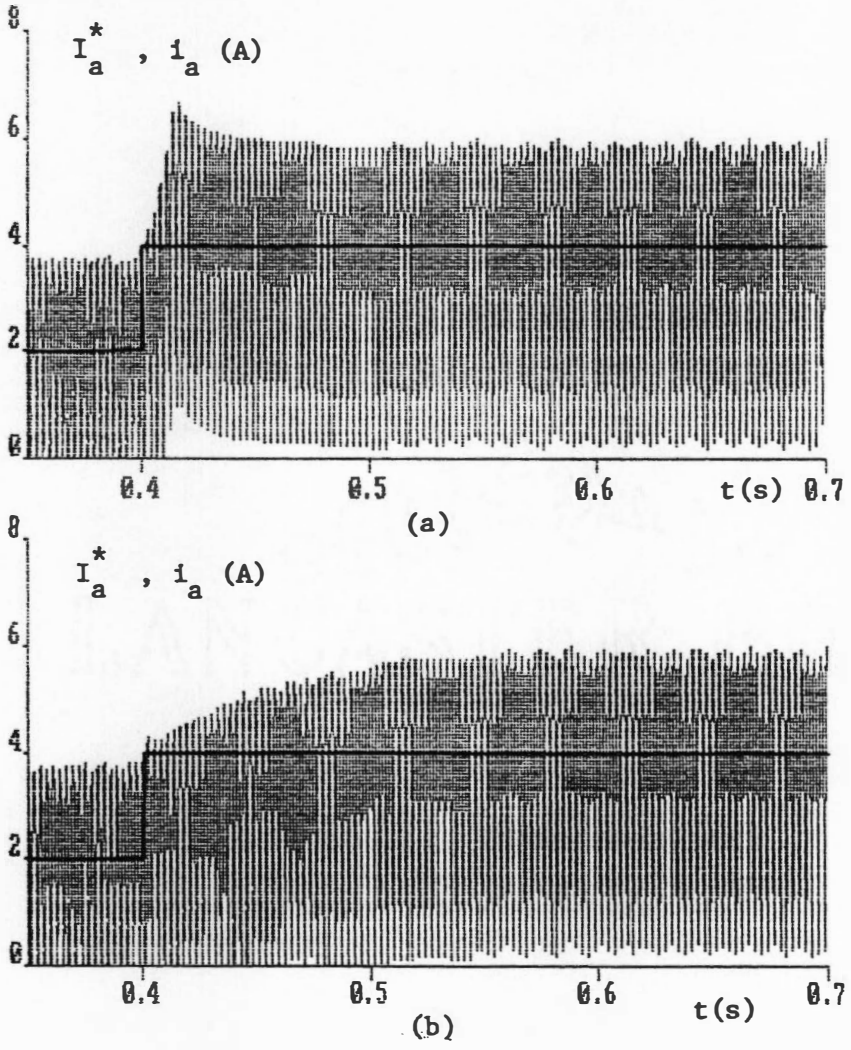


Fig. 2.10 Current loop response of the fuzzy controller.

a) With $\Delta\alpha$ compensation.

b) Without $\Delta\alpha$ compensation.

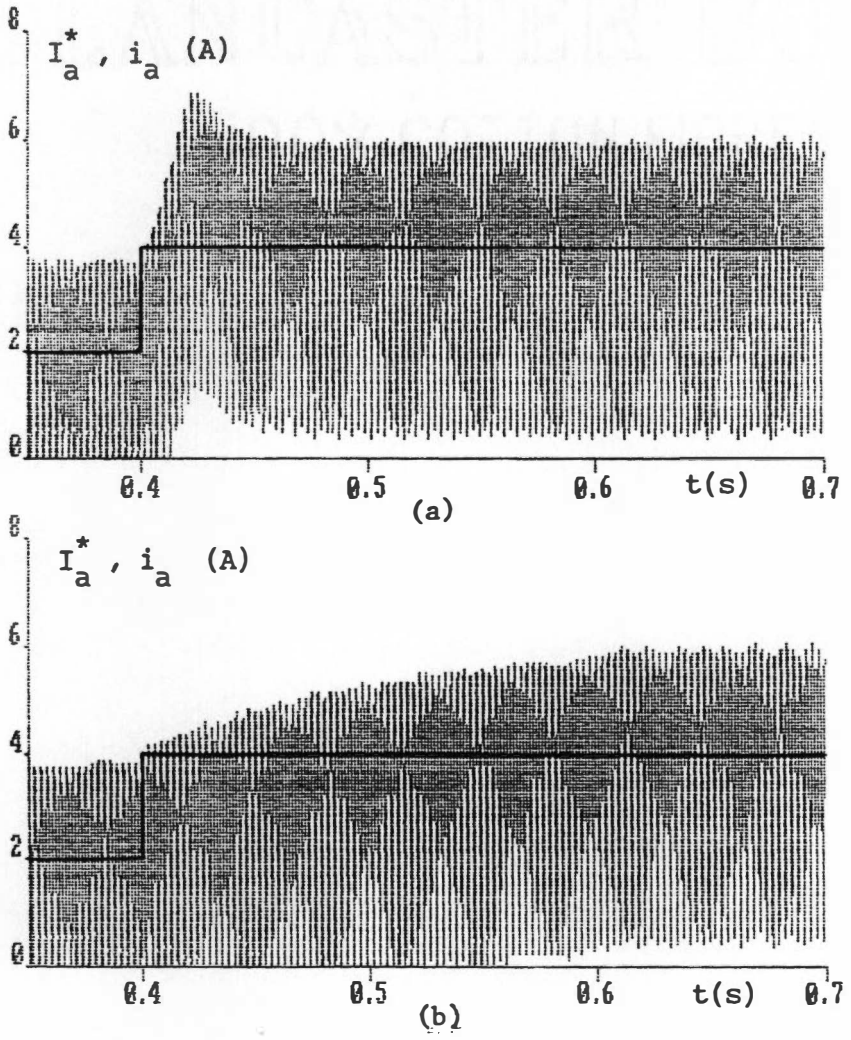


Fig. 2.11 Current loop response of the PI controller.

a) With $\Delta\alpha$ compensation.

b) Without $\Delta\alpha$ compensation.

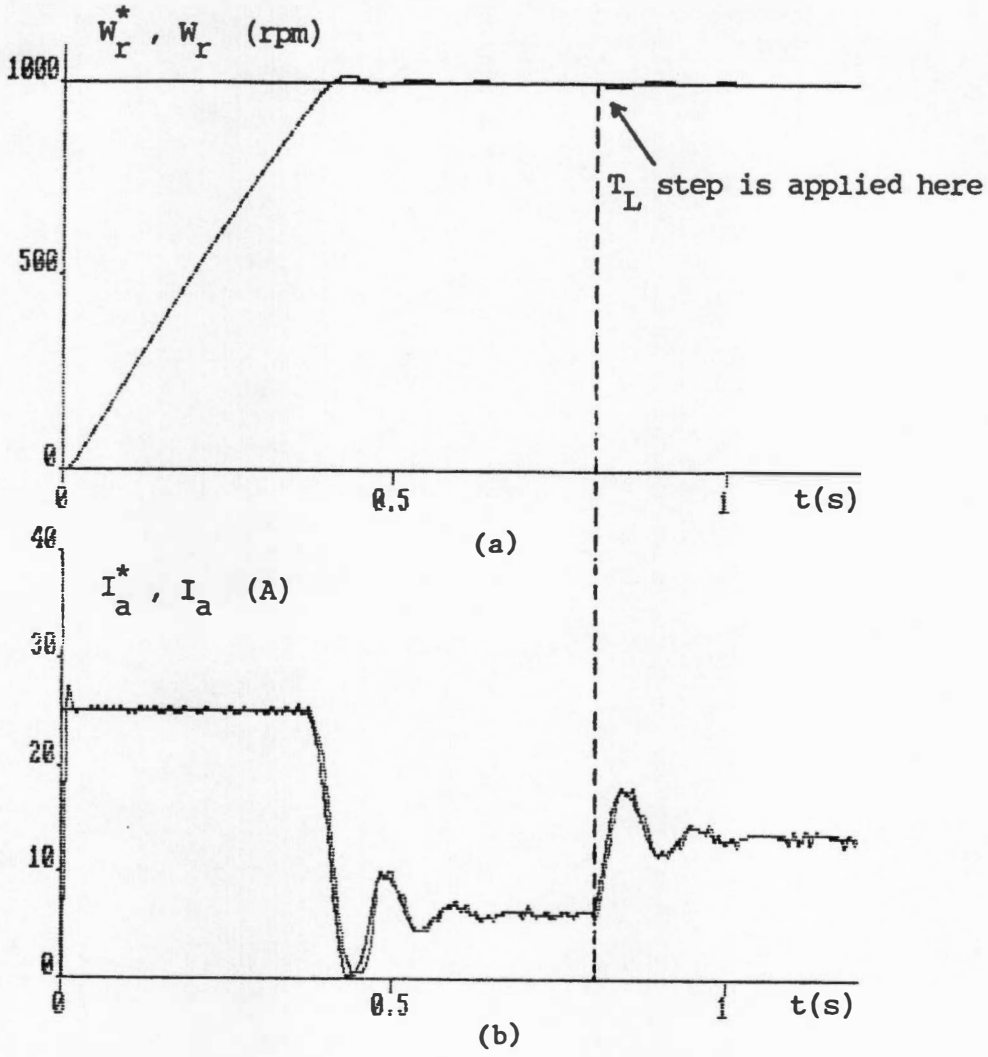


Fig. 2.12 Fuzzy control system response to a ω_r^* step and T_L step.

a) Speed. b) Armature currents.

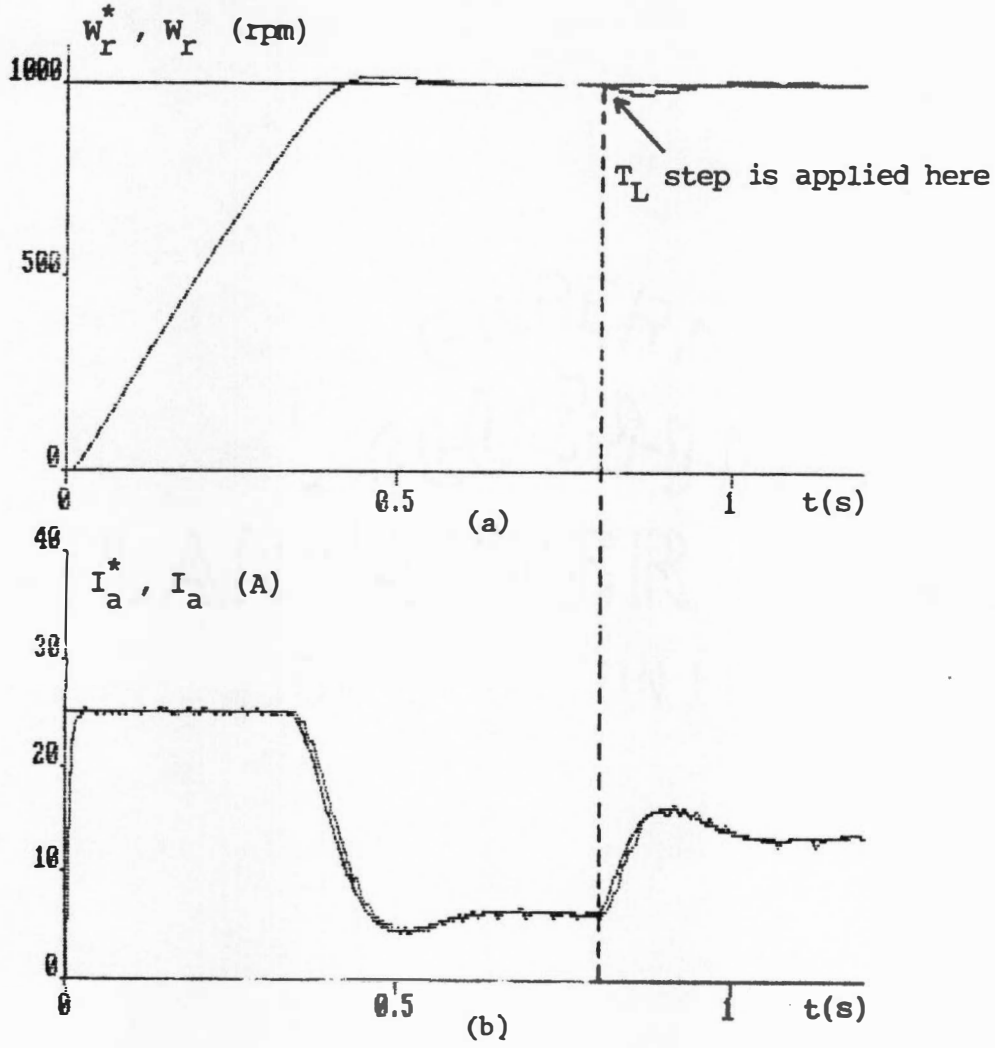


Fig. 2.13 PI control system response to a ω_r^* step and T_L step.

a) Speed. b) Armature currents.

corresponding system response under PI control in both loops. Table 2.4 summarizes the response improvement under fuzzy control.

Finally, the drive system was tested with ω_r^* step at the same condition as before but with four times the effective inertia load. Fig. 2.14 shows the response with fuzzy control, and Fig. 2.15 gives the response with PI control, for comparison. Although the major portion of the rise time occurs with the current loop saturated, some improvement in rise time and overshoot under fuzzy control is evident.

In conclusion, the simulation study demonstrated the successful application of fuzzy logic to a phase-controlled converter DC drive system. The proposed fuzzy $\Delta\alpha$ compensation scheme effectively linearizes converter characteristics at discontinuous conduction mode, and is simpler than the conventional look-up table method which requires a large memory. Fuzzy logic was also applied to the design of current and speed controllers, and the performance was compared with that of a conventional PI controlled system. The simulation study clearly indicated the superior performance of fuzzy control. The reason for superior performance of fuzzy controlled system is that basically it is adaptive in nature, and the controller is able to realize different control law for each input state (E and CE). The response of PI controlled system, on the other hand, is sensitive to model change that occurs with parameter variation. The main drawback in fuzzy control is the lack of a well defined design methodology, that results in a time consuming heuristic tuning process. Furthermore, a high speed microprocessor is needed to implement the control in real time. However, the control law is simpler to implement than expert system and neural network techniques.

Table 2.4 Performance comparison of fuzzy and PI controlled drive system.

	Fuzzy	PI
Rise time (s)	0.410	0.425
Overshoot (rpm)	13	16
Speed drop with T_L (rpm)	13	25
Recovery time with T_L(s)	0.09	0.20

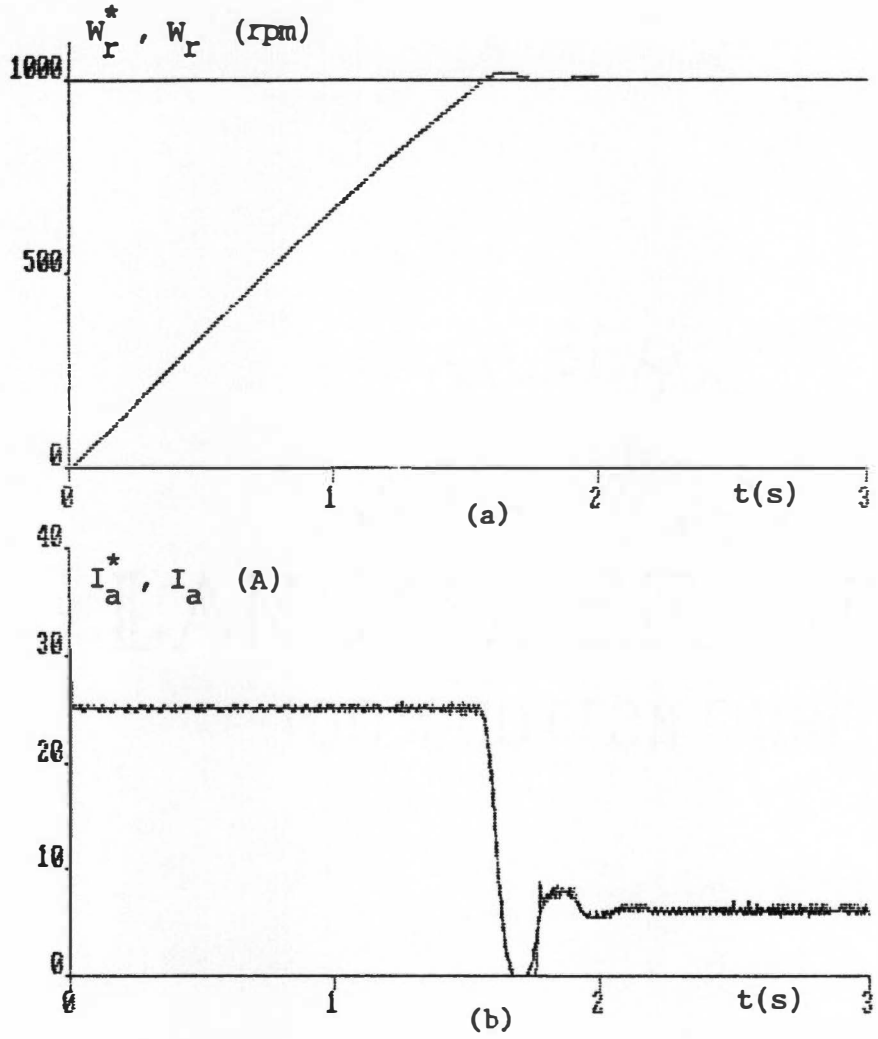


Fig. 2.14 Fuzzy control system response to a ω_r^* step with a new inertia of four times the original value.

a) Speed. b) Armature currents.

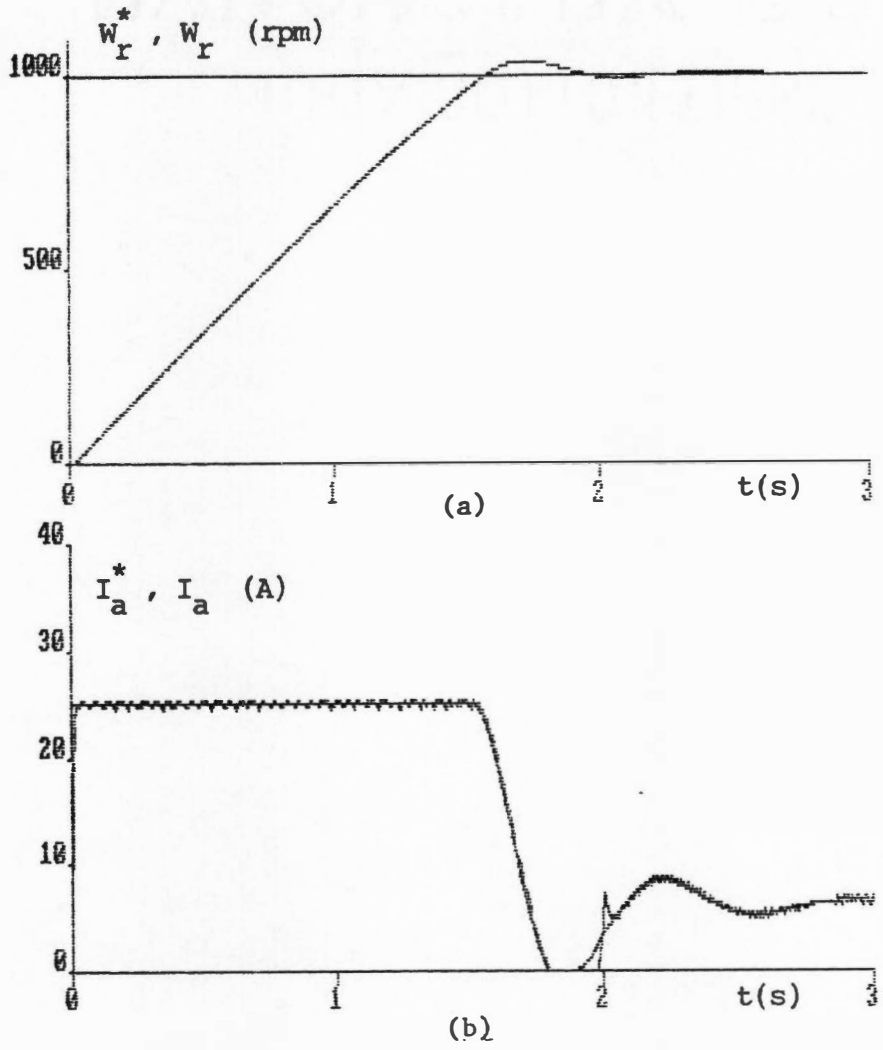


Fig. 2.15 PI control system response to a ω_r^* step with a new inertia of four times the original value.

a) Speed. b) Armature currents.

CHAPTER 3

LOSS MODELING OF CONVERTER INDUCTION MACHINE SYSTEMS

3.1 Introduction

Precise and reliable loss models for induction motor and converter systems are very important for performance prediction of variable speed drives. In particular, for the efficiency optimization study discussed in Chapter 4, it was realized that adequate converter and induction machine loss models were not available in the literature. In spite of the numerous studies that have been reported on the subject, in most cases the study focus on some particular aspects of interest and tend to neglect the overall picture, mainly because of the extreme complexity of the loss phenomena in induction machines. Traditionally, the machine electrical losses have been studied by using some variation of the per-phase equivalent circuit of Fig. 3.1, because the losses become primarily important at steady state condition. On the other hand, for dynamic studies, the synchronously rotating frame D^e-Q^e model of Fig. 3.2 is normally used, but in this case the losses are not properly represented. Even under sinusoidal supply, a precise evaluation of machine losses is not straightforward, but it becomes much more complex when the machine is fed by a inverter, due to the wide harmonic spectrum present in the impressed voltage. Nevertheless, in order to predict the performance of the fuzzy logic based efficiency

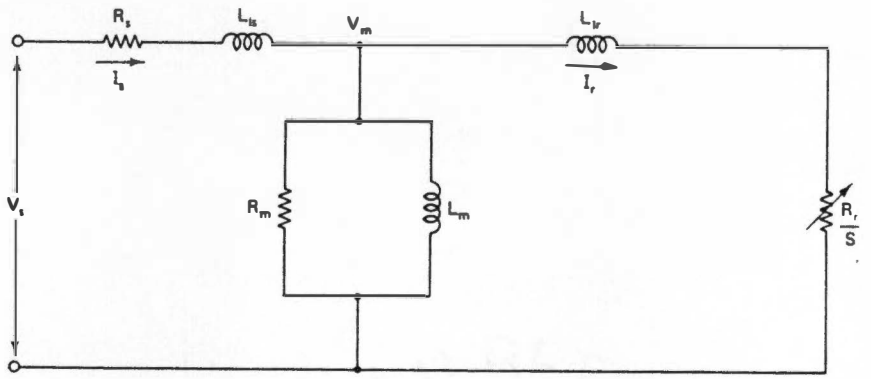


Fig. 3.1 Per-phase equivalent circuit.

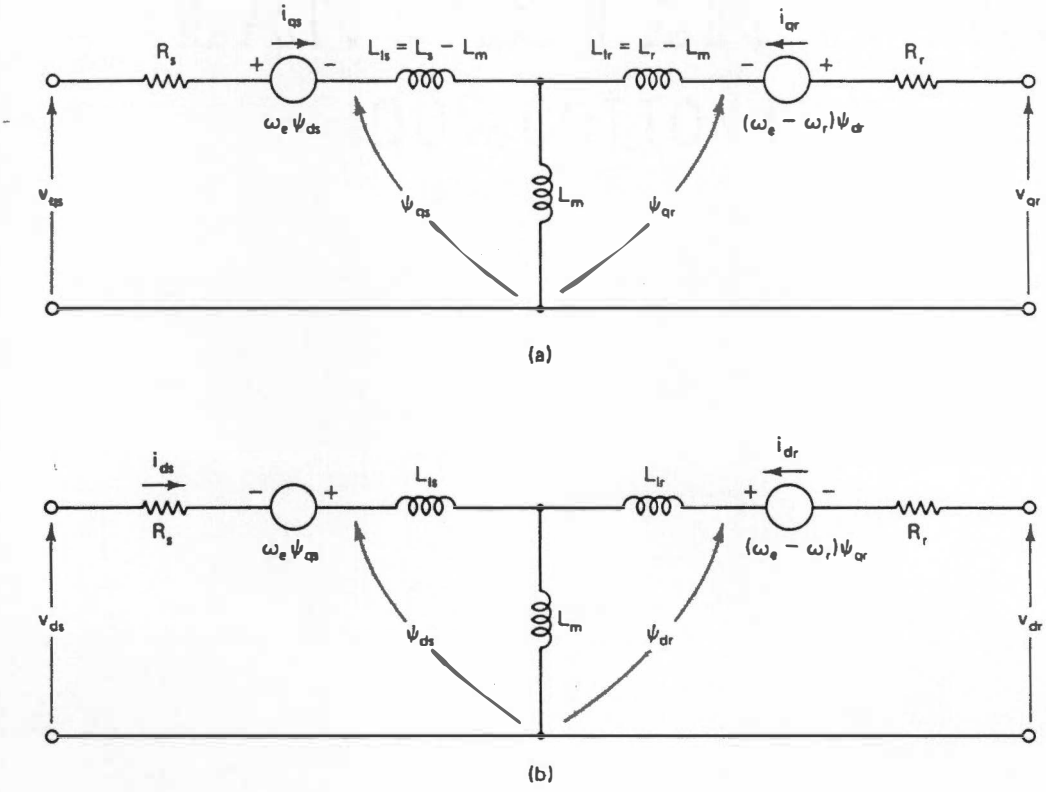


Fig. 3.2 Standard D^e -Q^e equivalent circuits in synchronous frame.

a) Q^e - axis circuit. b) D^e - axis circuit.

optimization control discussed in Chapter 4, it was necessary to derive a machine model capable of representing both the dynamic behavior and the loss characteristics of the induction machine, as well as a loss model for the converter system.

Loss modeling of induction motor has received wide attention in the literature over a number of years. Neglecting the effects of space harmonics, the machine parameters become dependent of time harmonic frequencies impressed by the inverter. Klingshirn and Jordan [30] applied superposition principle to calculate harmonic currents and the corresponding losses with per-phase equivalent circuit, where the rotor resistance and leakage inductance for each harmonic were corrected for deep bar (skin) effect. Kawagishi et al. [31] were able to verify experimentally the frequency dependency of parameter, by using a high frequency power supply, and validate some of the theoretical predictions. Honsinger [25] systematically studied the losses for a six-step inverter-fed machine and propose harmonic per-phase equivalent circuit. Since both core loss and stray load loss are basically due to hysteresis and eddy current effects, he proposed representation of stray loss by frequency dependent resistance in parallel with leakage inductance in the equivalent circuit. More recently, Udayagiri and Lipo [27] proposed a new simulation model that incorporates core loss but neglects the skin effect and leakage flux induced core loss, thereby underestimating the total loss.

In high frequency PWM inverter system, both conduction loss and switching loss are important in the loss model. The switching loss were discussed analytically by McMurray [32,] where the effects of both turn-on and turn-off snubbers were considered. Jovanovich et al. [33] made experimental evaluation of switching characteristics and

losses for a number of power devices with different base drives and load conditions. Ikeda et al. [34] have proposed loss modeling of PWM voltage-fed inverter and discussed the effect of carrier frequency on inverter losses. Circuit simulation programs, such as PSPICE that embed the detailed model of devices can give realistic lossy converter simulation. However, the drawbacks are that the losses remain somewhat transparent and can not be easily partitioned between the conduction and switching losses. Besides, such programs are not convenient for drive system simulation.

In the next sections, a unified loss model for the variable frequency drive system of Fig. 3.3 is discussed. The machine electrical losses, such as stator and rotor copper loss, core loss and stray loss, were considered for both fundamental and harmonic frequencies. Also considered were the skin effect on rotor resistance, temperature effect on both stator and rotor resistances, magnetizing inductance saturation, and friction and windage loss. All the above features were incorporated in a synchronous frame dynamic D^e-Q^e equivalent circuit. The converter system, consisting of a diode rectifier and PWM transistor inverter, was modeled accurately for conduction and switching losses. Validity of the models, in both steady-state and transient conditions, was verified by simulations. These results were first presented at the 1992 international conference on industrial electronics, control and instrumentation (IECON'92), [42].

3.2 Dynamic Lossy Model for the Induction Machine

The derivation of the dynamic lossy model is presented here. A discussion on machine losses is followed by the construction of a suitable harmonic equivalent circuit,

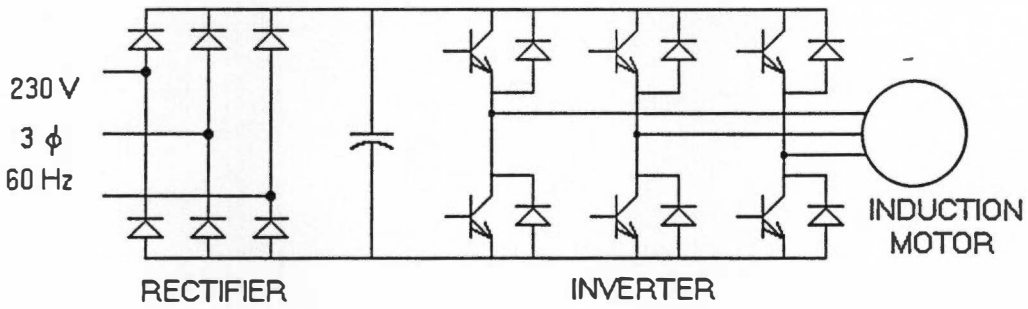


Fig. 3.3 Converter-machine system for variable speed drive.

and finally, by the derivation of the complete model through superposition principle.

3.2.1 Induction Machine Losses

Machine electrical and mechanical losses are examined here both qualitatively and quantitatively. The analysis follows a conventional partition of losses, encountered in the literature.

Copper Losses

Copper or Ohmic losses are caused by the non-ideal characteristics of the conductors employed in the stator and rotor windings, and have been modeled by

$$P_{cu} = m R I^2 \qquad (3.1)$$

where m is the number of phases, R is the winding resistance and I the rms current. Proper evaluation of these losses requires the consideration of temperature effect on winding resistance, as well as its frequency dependency, commonly referred to as skin effect.

The skin effect has been widely discussed in the literature. In inverter-fed machine, the skin effect due to fundamental slip frequency can be ignored, but for the harmonic frequencies the rotor appears almost stationary, and therefore, practically all the stator harmonic currents flow in the rotor creating dominant skin effect. The rotor resistance at harmonic frequency f_n is given approximately by [29]

$$R_{rn} = R_{rdc} \ (1 + c_1 \ d \ f_n^{0.5}) \tag{3.2}$$

where R_{rdc} = the dc resistance, d = bar depth, and c_1 is a constant that takes into account the bar material and shape. With a number of harmonic frequencies, the superposition principle can be applied approximately by assuming that machine parameters for all harmonic frequencies are identical to those computed at carrier frequency. For a PWM inverter with sinusoidal PWM or hysteresis-band current control, the carrier frequency and the near sidebands are most dominant.

Core Losses

Core losses consist of hysteresis and eddy-current losses, created by the time varying flux in the machine laminations. The eddy current loss may be expressed as

$$P_e = k_e f^2 \psi_m^2 \tag{3.3}$$

where ψ_m is the maximum mutual flux linkage, f the supply frequency and k_e a coefficient whose value depends on machine geometry and size, lamination thickness and resistivity of the iron. The hysteresis loss has been empirically modeled as

$$P_h = k_h f \psi_m^2 \tag{3.4}$$

where the coefficient k_h is mainly influenced by the magnetic properties of the iron and machine size. The stator core loss P_{cs} due to fundamental frequency mutual flux ψ_m can be expressed by

$$P_{cs} = k_h f \psi_m^2 + k_e f^2 \psi_m^2 \tag{3.5}$$

Under normal operation, the fundamental rotor frequency is a small fraction of the stator

frequency and the corresponding rotor core loss P_{cr} is given by

$$P_{cr} = k_h s f \psi_m^2 + k_e (sf)^2 \psi_m^2 \tag{3.6}$$

where f has been substituted for s f, s being the per unit slip. Equations 3.5 and 3.6 can be added and rearranged to get the total fundamental core loss as

$$P_c = P_{cs} + P_{cr} = \left[k_h \left(\frac{1+s}{f}\right) + k_e (1+s^2) \right] f^2 \psi_m^2 \tag{3.7}$$

As the mutual or air-gap flux linkage ψ_m is related to air-gap voltage V_m by

$$\psi_m = \sqrt{k_c} \frac{V_m}{f} \tag{3.8}$$

(3.7) can be rewritten as

$$P_c = k_c \left[k_h \left(\frac{1+s}{f} \right) + k_e (1+s^2) \right] V_m^2 \tag{3.9}$$

The equivalent core loss resistance R_m can then be derived as

$$R_m = \frac{1}{k_c \left[k_h \left(\frac{1+s}{f} \right) + k_e (1+s^2) \right]} \tag{3.10}$$

Assuming that the core losses due to mutual harmonic flux are governed by the same principles that control the fundamental losses, the coefficients k_h and k_e remain the same at harmonic frequency, and since harmonic slip $s_n \approx 1$, the equivalent core loss resistance R_{mn} at frequency f_n can be obtained from (3.10) as

$$R_{mn} = \frac{0.5}{k_c \left[\frac{k_h}{f_n} + k_e \right]} \tag{3.11}$$

Stray Losses

The stray losses by definition are the excess of the total losses actually occurring in a motor at a given load current, over the sum of the calculated copper losses, the no-load core loss, and the friction and windage loss, according to Alger et al. [35]. They used empirical equations to evaluate each individual loss component, what requires the knowledge of machine dimensions, type of core material, lamination thickness, winding geometry, etc. In this work, however, instead of evaluating stray loss individually, we treated them as a whole. The fundamental idea is that, the stray loss is essentially due to eddy current and hysteresis losses, induced by various types of leakage fluxes in the laminations and other structural parts of the machine. Therefore, the stray loss can be modeled in a way similar to that used for core loss modeling. The stator per phase stray loss at harmonic frequency f_n can be given as

$$P_{sln} = k_{sln}\left[\frac{k_h}{f_n} + k_e\right]V_{sln}^2 \qquad (3.12)$$

where V_{sln} is the voltage across the stator leakage inductance and k_{sln} the stray loss constant. Therefore, the stray loss can be represented by an equivalent resistance R_{sln} in parallel with the leakage inductance as

$$R_{sln} = \frac{1}{k_{sln}\left[\frac{k_h}{f_n} + k_e\right]} \qquad (3.13)$$

A similar expression was derived for rotor harmonic stray loss. The stray loss due to fundamental current is essentially concentrated in the stator, and an equation similar to (3.13) could also be used. However, the fundamental stray loss was represented by a

resistance in series with the stator leakage reactance X_{ls}, for reasons that will become clear later. The fundamental voltage drop V_{sl1} across the leakage reactance X_{ls} can be expressed as (2π f L_{ls} I_{s1}), where I_{s1} is the fundamental stator current. It can be substituted into (3.12) to derive fundamental per-phase stray loss P_{sl1} as

$$P_{sl1} = k_{sl1} [k_h f + k_e f^2] I_{s1}^2 = R_{sl1} I_{s1}^2 \qquad (3.14)$$

where R_{sl1} is the equivalent series resistance. From this expression, R_{sl1} is given as

$$R_{sl1} = k_{sl1} [k_h f + k_e f^2] \qquad (3.15)$$

Friction and Windage Losses

The friction and windage losses are essentially a function of motor speed ω_r and does not depend on the type of power supply. It can be expressed as

$$P_{fw} = k_{fw} \omega_r^3 \qquad (3.16)$$

3.2.2 Temperature and Saturation Effects

The preceding discussion on machine losses did not include the effects of temperature on winding resistances, nor the saturation effects on machine inductances. These effects will now be considered.

Temperature Effects

Both stator and rotor resistances increase with temperature. The stator temperature can be monitored and approximate correction factor can be applied, but there is no easy way to measure or estimate the rotor temperature. Precise prediction of temperature in each part of the machine requires detailed dynamic thermal model that depends on machine geometry, material characteristics, cooling effects, etc, and is extremely difficult to estimate. The machine transient thermal response can be given approximately by a first order model, where the temperature rise ΔT is expressed as

$$\Delta T = \frac{P_{tl}}{\theta (1 + \tau s)} \tag{3.17}$$

where P_{tl} is the total machine loss, θ is the steady-state thermal resistance, and τ is the thermal time constant. The θ and τ parameters can be estimated approximately by experimentation. Both rotor and stator resistances were corrected for temperature effects by using the well known formula

$$R_{T2} = R_{T1} (1 + \alpha_{T1} (T_2 - T_1)) \tag{3.18}$$

where α_{T1} = temperature coefficient (usually at T_1=25 °C), and $\Delta T = (T_2 - T_1)$. The temperature corrected resistances were then used to calculate fundamental and harmonic copper losses. For harmonic rotor losses, the skin effect was superimposed on the temperature effect.

Saturation Effects

Although saturation is strictly present in both leakage and magnetizing inductances, it was ignored in the former, and represented in the magnetizing inductance L_m by a piece-wise linear function of magnetizing current I_m as

$$\begin{aligned} L_m &= L_{mo} && , if\ I_m \le I_{mo} \\ &= L_{mo} - m\,(I_m - I_{mo}) && , if\ I_m > I_{mo} \end{aligned} \qquad (3.19)$$

where L_{mo} is the unsaturated inductance and I_m is the magnetizing current at the start of saturation. The saturation coefficient m was selected to best fit the actual saturation curve of the machine.

3.2.3 Per-phase Harmonic Equivalent Circuit

The effects of time harmonics have been traditionally investigated by solving the per-phase equivalent circuit [25] shown in Fig. 3.4(a), where the harmonic stray losses are represented by shunt resistances (R_{ssn} and R_{rsn}). For each harmonic component, the circuit is solved, and superposition principle is applied to get the overall harmonic effect [25]. In this way, the frequency dependence of machine parameters can be taken into account precisely. The following simplifying assumptions can be made at this point:

- For sinusoidal PWM or hysteresis-band current-controlled inverter, only the carrier frequency is considered for computation of frequency dependent parameters, and the resulting circuit can be used to compute the effect of all the harmonics with little loss of precision.
- The harmonic frequencies are sufficiently high such that the harmonic slip s_n is essentially one.

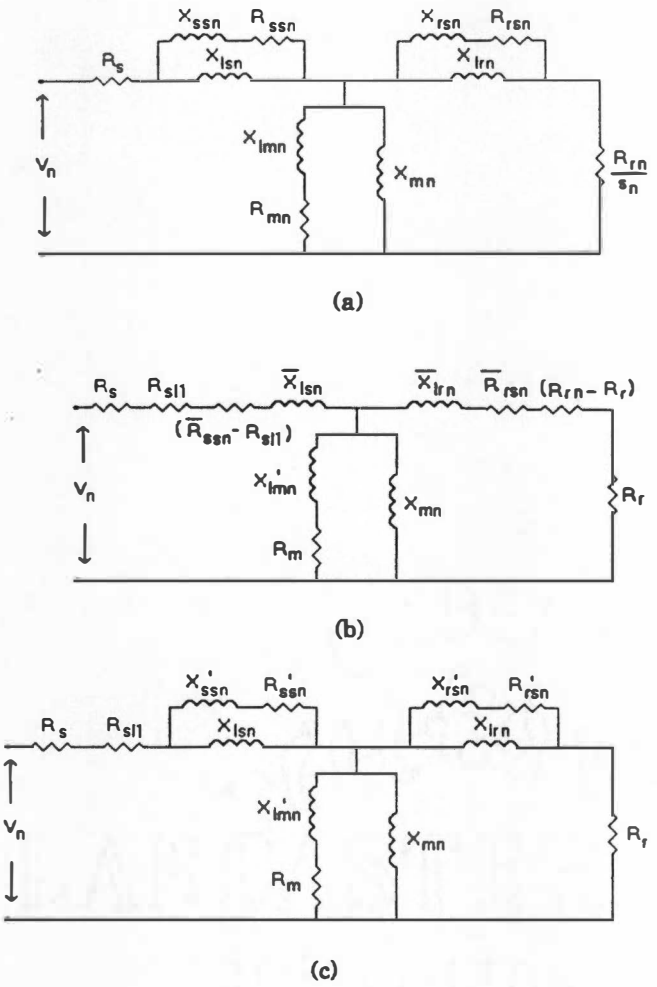


Fig. 3.4 Per phase harmonic equivalent circuit of the induction motor.

a) Generic circuit for harmonic of order n.

b) "Series" equivalent form of circuit (a).

c) Modified "shunt"equivalent form of circuit (b).

By applying these assumptions, Fig 3.4(a) was converted to the series equivalent form of Fig. 3.4(b). The barred parameters are simply the series equivalents of the corresponding original parameters. For example, the series equivalent stator stray loss resistance can be expressed as

$$\bar{R}_{ssn} = \frac{R_{ssn} X_{ssn}^2}{R_{ssn}^2 + (X_{ssn} + X_{lsn})^2} \cong \frac{R_{ssn} X_{ssn}^2}{R_{ssn}^2 + X_{lsn}^2} \tag{3.20}$$

since the secondary leakage reactance X_{ssn} is very small. The harmonic rotor resistance R_{rn} is shown split into fundamental rotor resistance R_r and $(R_{rn}-R_r)$. Similarly, R_{ssn} is shown as the sum of the fundamental frequency stray loss resistance R_{sl1} and $(R_{ssn}-R_{sl1})$ The harmonic core loss resistance R_{mn} was substituted in Fig. 3.4(b) by a series combination of the fundamental core loss resistance R_m and a modified secondary magnetizing reactance X_{lmn}', so as to ensure constancy of harmonic core loss P_{cln}. From Fig. 3.4(a), the harmonic core loss P_{cln} (neglecting small X_{lmn}) is given as

$$P_{cln} = \frac{3 V_{mn}^2}{R_{mn}} \tag{3.21}$$

where V_{mn} is the rms harmonic airgap voltage. From Fig. 3.4(b), P_{cln} is given by

$$P_{cln} = \frac{3 V_{mn}^2 R_m}{R_m^2 + X_{lmn}'^2} \tag{3.22}$$

In order to keep the harmonic core loss invariant, the two equations must yield the same result. By equating the two expressions, the modified secondary magnetizing reactance was derived as follows:

$$X_{lmn}' = \sqrt{R_{mn} R_m - R_m^2} \tag{3.23}$$

The circuit of Fig. 3.4(b) was next transformed into the modified "shunt" form of Fig. 3.4(c). The final values of the harmonic stray loss resistances R_{ssn}' and R_{rsn}' were obtained by equating the corresponding resistive terms in Figs. 3.4(b) and 3.4(c). Neglecting the small X_{rsn}' the following expression was derived:

$$\frac{R_{rsn}' X_{lrn}}{R_{rsn}'^{2} + X_{lrn}^2} = R_{rle} \tag{3.24}$$

where $R_{rle} = R_{rsn} + (R_{rn} - R_r)$. Solving for R_{rsn}',

$$R_{rsn}' = \frac{X_{lrn}^2 \pm \sqrt{\left(\frac{X_{lrn}^2}{R_{rle}}\right)^2 - 4 X_{lrn}^2}}{2} \tag{3.25}$$

For most practical drives, $R_{rsn} > X_{lrn}$ and therefore, the plus sign is considered in the above equation. With a similar procedure, the expression for R_{ssn}' was also derived. In practice, the value of R_{sll} is very small compared to R_{ssn}. Therefore, R_{ssn}' could be taken equal to R_{ssn}. Note that R_{rsn}' represents not only the rotor harmonic stray loss, but also the additional harmonic copper loss due to skin effect.

3.2.4 Synchronous Frame Lossy Equivalent Circuits

The per-phase equivalent circuit derived in Fig. 3.4(c) is only valid for steady-state operation, and can not be used for dynamic performance study. Usually, synchronously rotating frame D^e-Q^e equivalent circuits are used for dynamic study. The standard D^e-Q^e equivalent circuits can not directly incorporate a core loss resistor in parallel with magnetizing inductance because DC current (equivalent fundamental frequency current)

will not flow through it. In this section, it is described how these circuits were modified to incorporate the core loss resistor, and then how the harmonic equivalent circuits were superimposed to them to derive the unified lossy equivalent circuits.

D^e-Q^e Equivalent Circuits with Core Loss Resistor

The stationary frame D^s-Q^s equivalent circuits can easily incorporate core loss resistance in parallel with magnetizing inductance, as shown in Fig. 3.5. With this modifications the following equations can be readily written:

$$R_s\begin{pmatrix} i_{qs}^s \\ i_{ds}^s \end{pmatrix} + L_{ls}\frac{d}{dt}\begin{pmatrix} i_{qs}^s \\ i_{ds}^s \end{pmatrix} + R_m\begin{pmatrix} i_{qrm}^s \\ i_{drm}^s \end{pmatrix} = \begin{pmatrix} v_{qs}^s \\ v_{ds}^s \end{pmatrix} \tag{3.26}$$

$$-L_m\frac{d}{dt}\begin{pmatrix} i_{qm}^s \\ i_{dm}^s \end{pmatrix} + R_m\begin{pmatrix} i_{qrm}^s \\ i_{drm}^s \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \end{pmatrix} \tag{3.27}$$

$$-\omega_r\begin{pmatrix} \Psi_{dr}^s \\ \Psi_{qr}^s \end{pmatrix} + R_r\begin{pmatrix} i_{qr}^s \\ i_{dr}^s \end{pmatrix} + L_{lr}\frac{d}{dt}\begin{pmatrix} i_{qr}^s \\ i_{dr}^s \end{pmatrix} + L_m\frac{d}{dt}\begin{pmatrix} i_{qm}^s \\ i_{dm}^s \end{pmatrix} = \begin{pmatrix} v_{qr}^s \\ v_{dr}^s \end{pmatrix} \tag{3.28}$$

Synchronous frame quantities are related to their stationary frame counterparts by a transformation matrix T, defined as

$$T = \begin{bmatrix} \cos\omega_e t & \sin\omega_e t \\ -\sin\omega_e t & \cos\omega_e t \end{bmatrix} \tag{3.29}$$

where ω_e is the synchronous frequency, $\cos\omega_e t$ and $\sin\omega_e t$ are the unit vectors. For the stator currents, the relationship can be expressed as

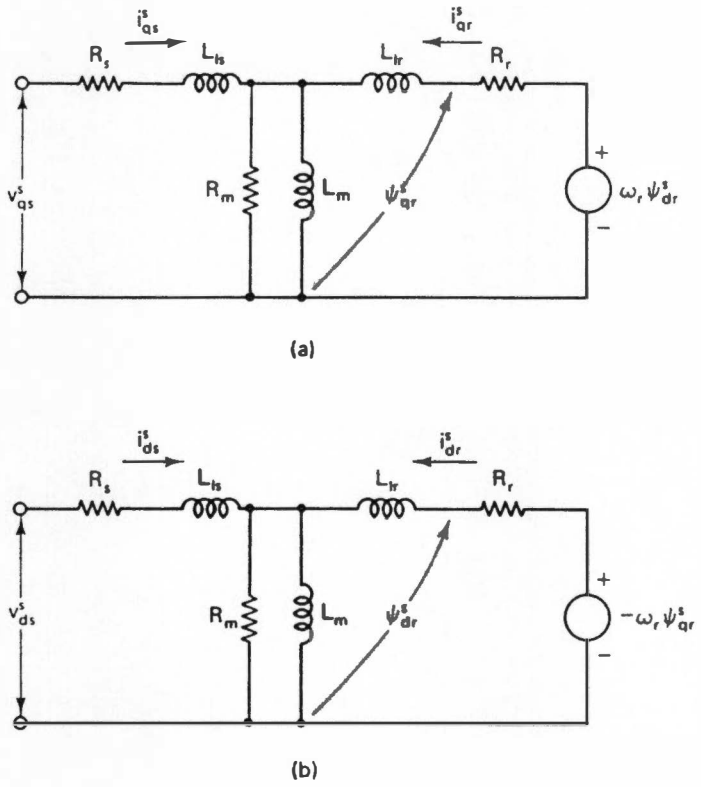


Fig. 3.5 D^s-Q^s equivalent circuit with core loss resistance.

$$\begin{pmatrix} i_{qs}^{s} \\ i_{ds}^{s} \end{pmatrix} = T \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \tag{3.30}$$

Substitution of (3.30) and similar relations for the remaining variables into (3.26) results in

$$R_s T \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + L_{ls} \frac{d}{dt}\left[T \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \right] + R_m T \begin{pmatrix} i_{qrm} \\ i_{drm} \end{pmatrix} = T \begin{pmatrix} v_{qs} \\ v_{ds} \end{pmatrix} \tag{3.31}$$

Pre-multiplying (3.31) by the inverse transformation T^{-1} yields

$$R_s \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + L_{ls} T^{-1} \frac{d}{dt}\left[T \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \right] + R_m \begin{pmatrix} i_{qrm} \\ i_{drm} \end{pmatrix} = \begin{pmatrix} v_{qs} \\ v_{ds} \end{pmatrix} \tag{3.32}$$

As T is a function of time, the product rule must be used to evaluate the derivative, and the second term can be rewritten as

$$\begin{aligned} L_{ls} T^{-1} \frac{d}{dt}\left[T \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \right] &= L_{ls} \left\{ T^{-1} \frac{dT}{dt} \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + T^{-1} T \frac{d}{dt} \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \right\} \\ &= L_{ls} \left\{ \begin{pmatrix} 0 & \omega_e \\ -\omega_e & 0 \end{pmatrix} \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + \frac{d}{dt} \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \right\} \\ &= L_{ls} \omega_e \begin{pmatrix} i_{ds} \\ -i_{qs} \end{pmatrix} + L_{ls} \frac{d}{dt} \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} \end{aligned} \tag{3.33}$$

Substitution of the last form of (3.33) into (3.32) produces

$$R_s\begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + \omega_e \begin{pmatrix} \psi_{dls} \\ -\psi_{qls} \end{pmatrix} + L_{ls}\frac{d}{dt}\begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix} + R_m\begin{pmatrix} i_{qrm} \\ i_{drm} \end{pmatrix} = \begin{pmatrix} v_{qs} \\ v_{ds} \end{pmatrix} \qquad (3.34)$$

where $\psi_{dls} = L_{ls}\, i_{ds}$ and $\psi_{qls} = L_{ls}\, i_{qs}$ are the stator leakage fluxes, and the new terms $\omega_e\, \psi_{dls}$ and $-\omega_e \psi_{qls}$ represent the speed voltages due to rotation of the reference axes. By applying the same methodology to (3.27) and (3.28), their synchronous frame counterparts can be obtained as

$$\omega_e\begin{pmatrix} \psi_{dm} \\ \psi_{qm} \end{pmatrix} + L_m\frac{d}{dt}\begin{pmatrix} i_{qm} \\ i_{dm} \end{pmatrix} - R_m\begin{pmatrix} i_{qrm} \\ i_{drm} \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \end{pmatrix} \qquad (3.35)$$

$$(\omega_e - \omega_r)\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} + R_r\begin{pmatrix} i_{qr} \\ i_{dr} \end{pmatrix} + L_{lr}\frac{d}{dt}\begin{pmatrix} i_{qr} \\ i_{dr} \end{pmatrix} + L_m\frac{d}{dt}\begin{pmatrix} i_{qm} \\ i_{dm} \end{pmatrix} = \begin{pmatrix} v_{qr} \\ v_{dr} \end{pmatrix} \qquad (3.36)$$

Equation 3.34 through 3.36 can be represented by the equivalent circuits of Fig.3.6, where $\psi_{dm} = L_m i_{dm}$ and $\psi_{qm} = L_m i_{qm}$ are the air-gap fluxes, $\psi_{dr} = \psi_{dm} + L_{lr} i_{dr}$ and $\psi_{qr} = \psi_{qm} + L_{lr} i_{qr}$ are the rotor fluxes.

D^e-Q^e Equivalent Circuits with Core Loss and Harmonic Loss

Although the harmonic equivalent circuit of Fig. 3.4(c) is a per phase stationary frame circuit, it can be used in the synchronously rotating frame as well, with little loss of precision, because the axes rotation will have the effect of adding (or subtracting) the fundamental frequency to the particular harmonic frequency. Consequently, under the assumption $f_n >> f$, the effect of axes rotation is negligible. The circuit of Fig. 3.4(c) was therefore, superimposed on the synchronously rotating frame D^e-Q^e equivalent circuits with core loss resistance of Fig. 3.6, resulting in the complete lossy dynamic D^e-Q^e

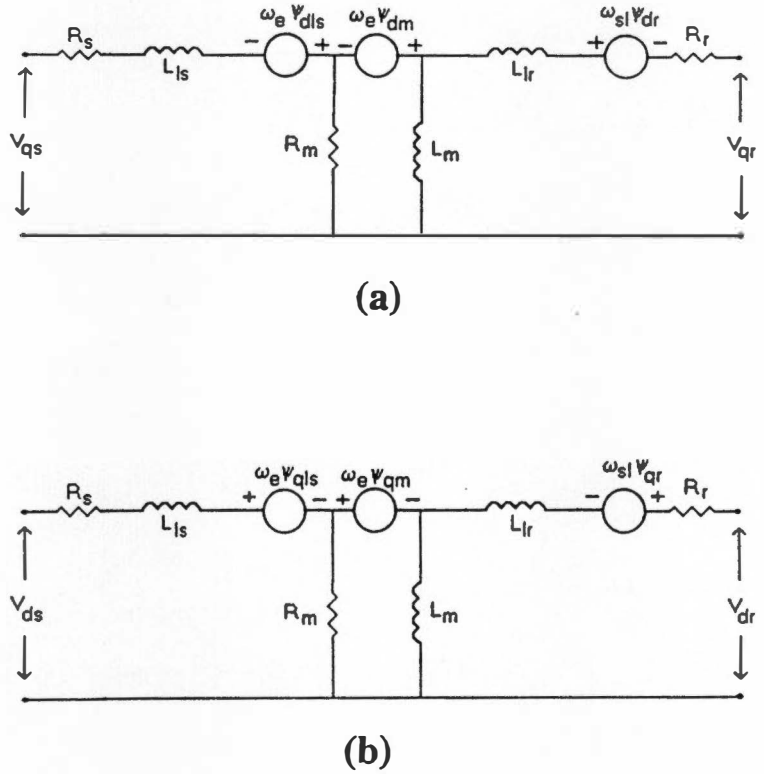
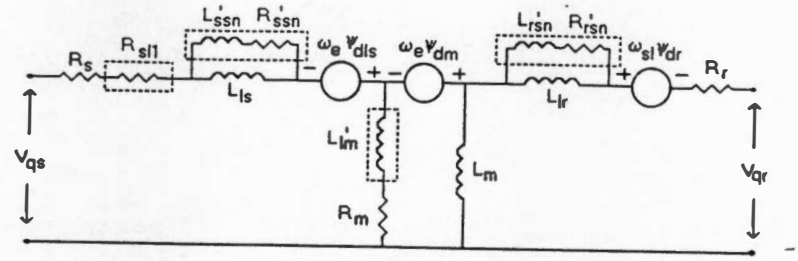


Fig. 3.6 D^e-Q^e equivalent circuit with core loss resistance.

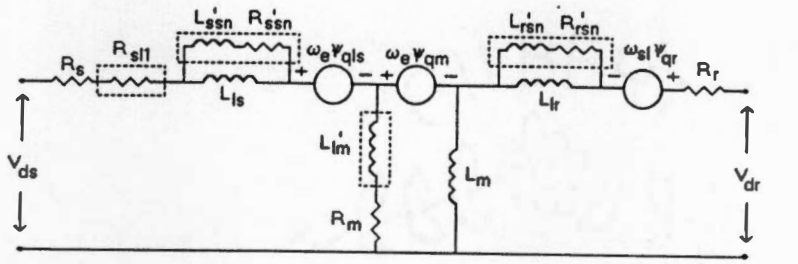
equivalent circuits shown in Fig. 3.7. The model represents true physical behavior of the machine, and can be used for evaluation of fundamental and harmonic losses, as well as for dynamic studies. Since at steady-state conditions all fundamental variables appear as DC quantities, the fundamental current will not flow through R'_{ssn} and R'_{rsn}, i.e., the harmonic stray loss shunt terms have no effect on fundamental losses. Similarly, the presence of the secondary magnetizing inductance L'_{lm} has no effect on fundamental core loss. For harmonic losses, the circuits of Fig. 3.7 give almost the same result as that of Fig. 3.4(c). This follows from the fact that the harmonic fluxes are very small, and therefore, the corresponding harmonic voltages present in the dependent voltage sources are negligible, i.e., the terms such as $\omega_e \psi_{dm}$ essentially represent fundamental counter emf. During transient operations, some amount of fundamental current will flow through the shunt stray loss resistances, and this is theoretically coherent, because a change in fundamental leakage flux induces extra stray losses both in the stator and the rotor. During the transient, the effective rotor resistance seen by the fundamental current increases and the effective inductance decreases, and this is also consistent with the theory, because the skin effect is present during sinusoidal transients.

3.2.5 Lossy Model Parameter Computation

In order to use the equivalent circuits of Fig. 3.7, it was necessary to evaluate the parameters for the particular machine. The fundamental frequency parameters can be easily obtained from standard tests, but harmonic frequency parameters are much more difficult to obtain experimentally [28]. On the other hand, accurate theoretical prediction



a)



b)

3.7 Lossy D^e-Q^e equivalent circuits in synchronously rotating reference frame.

a) Q^e-axis circuit b) D^e-axis circuit

of harmonic frequency parameters also constitutes a formidable task. In this work, the standard fundamental frequency parameters were taken from literature [39], and the remaining parameters were estimated from practical considerations [31], as discussed below.

The fundamental stray load loss P_{sl1} at rated conditions amount to a few percent (typically 1% to 3%) of the rated machine power P_o [35]. Let ε_{sl} be the selected (or measured) percentage for a particular machine with a rated current I_s. The stray loss resistance R_{slb}, at base frequency f_b is given by

$$R_{slb} = \frac{\epsilon_{sl} P_o}{3 I_s^2} \tag{3.37}$$

At any other fundamental frequency f, the fundamental stray loss resistance of (3.15) can be rewritten as a function of R_{slb}

$$R_{sl1'} = R_{slb}\left(\frac{f}{f_b}\right)\left(\frac{1+\gamma_1 f}{1+\gamma_1 f_b}\right) \tag{3.38}$$

where $\gamma_1 = K_e/k_h$.

Although the core loss resistance R_m at base 60 Hz frequency can be easily computed from standard machine tests, the estimation of eddy current coefficient k_e and hysteresis loss coefficient k_h requires the breakdown of total core loss P_c into its eddy current P_e and hysteresis P_h components. Usually, the eddy current loss P_e amounts to 20% - 50% of total core loss. Let ε_{ec} be the assumed P_e/P_c ratio. The corresponding relation between eddy current and hysteresis losses becomes

$$P_e = \frac{\epsilon_{ec}}{1-\epsilon_{ec}} P_h \tag{3.39}$$

Substituting (3.3) and (3.4) into (3.39) leads to

$$k_e = \frac{k_h}{f} \frac{\epsilon_{ec}}{1-\epsilon_{ec}} \tag{3.40}$$

At no load, the per unit slip s is practically zero, and (3.10) can be rewritten as

$$\frac{k_h}{f} + k_e = \frac{(2\pi)^2}{R_m} \tag{3.41}$$

where $(2\pi)^2$ has been substituted for k_c. Finally, solving (3.40) and (3.41) results in

$$k_e = \frac{(2\pi)^2 \epsilon_{ec}}{R_m} \tag{3.42}$$

$$k_h = \frac{(2\pi)^2 (1-\epsilon_{ec}) f}{R_m} \tag{3.43}$$

The harmonic frequency stray loss resistances will now be discussed. Kawagishi et al. [31] have indicated that the efficiency of an induction machine fed by a 5 KHz sinusoidal PWM inverter is nearly 99% of that attainable under a 60 Hz AC supply. Their findings were also confirmed by Bausch el al. [59] for a much larger machine, operating from a hysteresis band PWM inverter. It can be assumed, therefore, that the harmonic stray losses P_{slh} of an induction motor fed by a 5 KHz PWM inverter is a percentage ϵ_{slh} ($\approx$1%) of the rated output power P_o. From Fig. 3.4(a) it is clear that, essentially all harmonic voltage will appear across the leakage reactances X_{lsn} and X_{lrn}, since they are much larger than the resistive terms R_s and R_r. Neglecting the small

secondary leakage reactance terms X_{ssn} and X_{rsn}, the total harmonic stray loss resistance R_{slh} can be obtained from

$$R_{slh} = \frac{3V_h^2}{\epsilon_{slh} P_o} \tag{3.44}$$

where V_h is total rms harmonic phase voltage. For a given PWM inverter, V_h can be obtained from simulation, or computed from the harmonic spectrum, if known. Distribution of R_{slh} into rotor and stator components is accomplished by assuming a partition proportional to the respective leakage inductances as

$$R_{rsb} = R_{slh} \frac{L_{lr}}{L_{lr} + L_{ls}} \tag{3.45}$$

$$R_{ssb} = R_{slh} \frac{L_{ls}}{L_{lr} + L_{ls}} \tag{3.46}$$

The actual harmonic stray loss resistances R_{ssn} and R_{rsn}, at any other carrier frequency f_c are obtained from (3.13) as

$$R_{rsn} = R_{ssb} \left(\frac{f_c}{f_{cb}} \right) \left(\frac{1 + \gamma_h f_{cb}}{1 + \gamma_h f_c} \right) \tag{3.47}$$

$$R_{rsn} = R_{rsb} \left(\frac{f_c}{f_{cb}} \right) \left(\frac{1 + \gamma_h f_{cb}}{1 + \gamma_h f_c} \right) \tag{3.48}$$

where $\gamma_h = k_{eh}/k_{hh}$ is assumed to be identical to γ_1, and f_{cb} is the base carrier frequency (5 KHz, in this discussion).

The secondary leakage inductances L_{ssn} and L_{rsn}, due to eddy current fluxes, are small compared to the conventional leakage inductances [25]. In the simulation study

discussed later, they were taken as 5% of the corresponding leakage inductances. Their presence in the equivalent circuit is mainly to ensure consistency of the dynamic behavior of the model. In fact, without them the equivalent circuit would become purely resistive at infinite frequency, whereas the actual induction motor becomes purely inductive.

3.3 Loss Modeling of Converter System

While for a diode rectifier only the conduction loss need to be considered, for a PWM inverter both switching loss and conduction loss are relevant.

3.3.1 Loss Modeling of Diode Rectifier

The three-phase diode rectifier with capacitor filter of Fig. 3.3 conducts discontinuously, and it can be shown that only two diodes (one in the upper group and one in the lower group) conduct at any instant. Fig. 3.8 shows the Thevenin equivalent circuit, where V_r is the ideal rectifier voltage profile, and the DC link inductance is $L_d = 2\ L_{sl}$, where L_{sl} is the per-phase source leakage inductance. The typical DC link voltages and current waves are shown in Fig.3.9. The diodes conduct a current pulse i_d when $V_r > (V_d + V_f)$, but conduction ceases sometime after $V_r < (V_d + V_f)$, due to L_d effect. The two conducting diodes can be modeled by an off-set voltage V_f in series with a nonlinear resistance R as indicated. From the conduction characteristics of a particular diode, the voltage drop equation can be derived by using the software TABLE-CURVE [36]. The software takes a set of (x,y) coordinate pairs as input and calculates the equation of the best fitting curve as given by

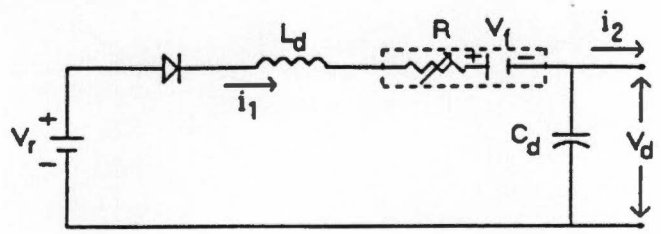


Fig. 3.8 Diode rectifier equivalent circuit.

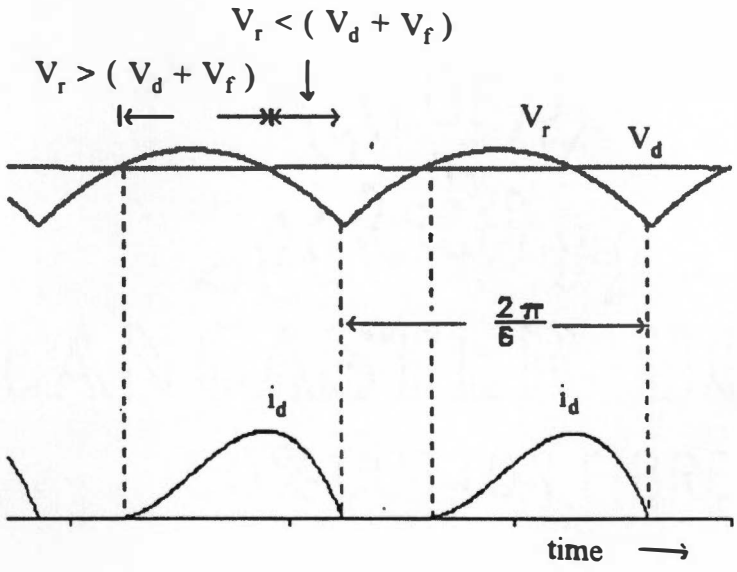


Fig. 3.9 Rectifier voltage and current waves

$$v_{dd} = v_{do} + K i_d^m \tag{3.49}$$

where v_{d0} is the offset voltage, and m the resistive drop exponent. The instantaneous conduction loss P_{ild} for the diode bridge can then be given as

$$P_{ild} = 2 v_{dd} i_d \tag{3.50}$$

Similarly, the instantaneous rectifier input power P_{iir} and output power P_{ior} can be expressed as

$$P_{iir} = V_r i_d \tag{3.51}$$

$$P_{ior} = V_d i_d \tag{3.52}$$

In the simulation program these expressions are integrated and averaged over a 60° interval to get the corresponding average values, required for efficiency computations.

3.3.2 Loss Modeling of PWM Inverter

For a PWM inverter, both conduction and switching losses should be considered. The conduction loss, in turn, is distributed between transistors and feedback diodes.

Conduction Loss

A typical inverter phase leg, with feedback diodes and polarized snubbers, is shown in Fig. 3.10(a). The conduction loss equivalent circuit is shown in Fig. 3.10(b), where each conduction drop has been modeled by a resistance in series with an offset voltage. Again, from the transistor saturation characteristics, the following linear voltage drop equation was derived with the help of TABLE-CURVE as shown in (3.53).

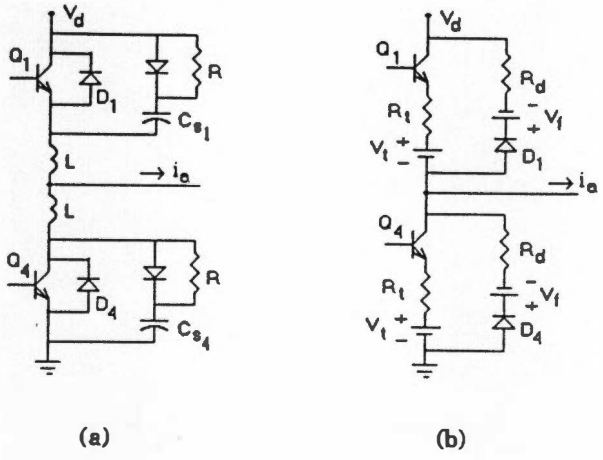


Fig. 3.10 (a) Transistor inverter phase leg.

(b) Conduction loss equivalent circuit.

$$v_{td} = v_{t0} + R_t i_c \tag{3.53}$$

For the feedback diodes, the voltage drop eqn. (3.48) is valid for the particular devices.

It is important to observe that, for the positive half cycle of phase current i_a, only transistor Q_1 or feedback diode D_4 conducts, whereas for the negative half cycle, Q_4 or D_1 conducts. Let G_A be the logic signal from the PWM control, such that, if G_A=1 the upper transistor Q_1 conducts and conversely, if G_A= 0, Q_4 conducts. The leg voltage with respect to "ground", v_{ag}, can be derived from the equivalent circuit as follows

$$\begin{aligned} v_{ag} &= V_d - v_{tda} \; ; \; if \; i_a>0 \; and \; G_A=1 \\ &= -v_{dda} \quad ; \; if \; i_a>0 \; and \; G_A=0 \\ &= V_d + v_{dda} \; ; \; if \; i_a<0 \; and \; G_A=1 \\ &= v_{dda} \quad ; \; if \; i_a<0 \; and \; G_A=0 \end{aligned} \tag{3.54}$$

Conversely, the instantaneous conduction loss in phase leg A can be expressed as

$$\begin{aligned} P_{cla} &= (V_d - v_{ag})\, i_a \quad ; \; if \; G_A=1 \\ &= -v_{ag}\, i_a \quad ; \; if \; G_A=0 \end{aligned} \tag{3.55}$$

Similar expressions can be written for the other two phase legs, and the total instantaneous conduction loss P_{icli} obtained as the sum of the three components.

Switching Losses

Proper computation of switching losses requires careful analysis of inverter operation, as illustrated in Fig. 3.11 by the PWM waves for phase leg a. Assume, for example, that i_a is positive and Q_1 is initially ON, such that current flows through it.

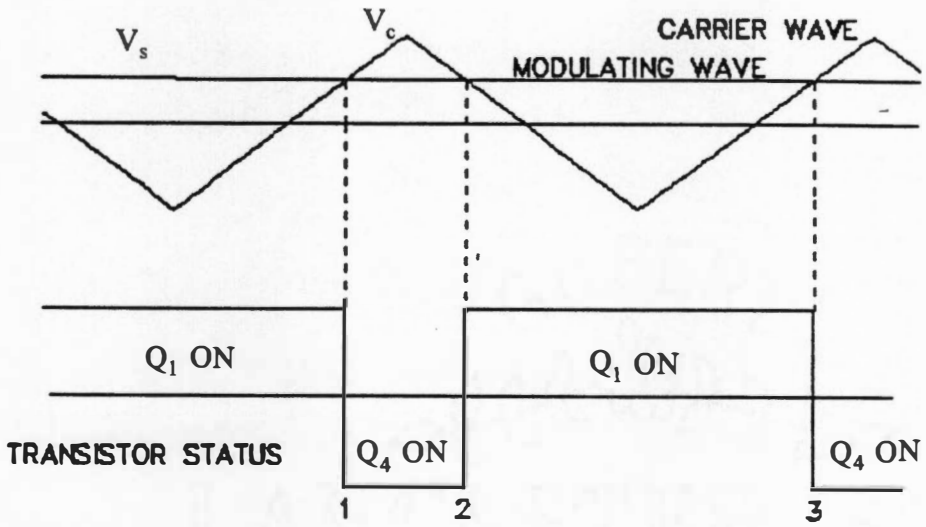


Fig. 3.11 Pulse width modulation waves for phase leg A.

At instant 1, Q_1 is turned OFF and Q_4 is turned ON, the energy stored in C_{s4} is dissipated across the series resistor, and Q_1 goes through a turn-off switching loss. However, as the current is now actually flowing through D_4, no turn-on loss occurs in Q_4. Similarly, at instant 2, when Q_1 is again turned ON, C_{s1} discharges and at the same time the device has a turn-on loss, but no loss occurs in Q_4, since it was not carrying current. Typical turn-on and turn-off switching waves are shown in Fig. 3.12. The energy loss during switchings can be related to the area of the product curve v_{CE} i_C (not shown). Summarizing, at every period of the carrier wave, the switching losses per phase leg are:

- Two discharges of capacitor snubbers;
- One turn-on loss;
- One turn-off loss.

Shunt Snubber Loss

After the qualitative discussion above, loss expressions will be derived. The energy loss associated with one snubber capacitor C_s discharge is

$$W_1 = \frac{1}{2} C_s V_d^2 \tag{3.56}$$

The total snubber energy loss of the inverter, in one fundamental period T, can be represented as

$$W_t = 3 N_s W_1 = \frac{3}{2} N_s C_s V_d^2 \tag{3.57}$$

where N_s is the number of switchings (on-and-off) of a phase leg in one fundamental period T. Consequently, the average power loss due to snubber discharge is

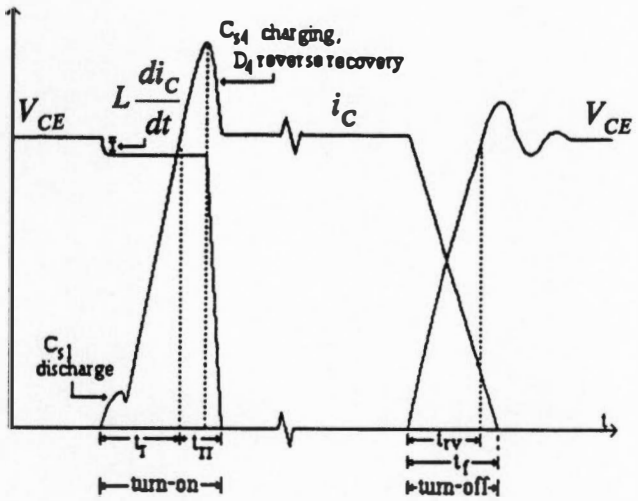


Fig. 3.12 Typical turn-on and turn-off switching waves for transistor Q_1.

$$P_s = \frac{3}{2}\frac{N_s C_s V_d^2}{T} \tag{3.58}$$

This loss can be represented by an equivalent shunt resistance R_s across the DC bus as

$$R_s = \frac{2}{3fN_s C_s} \tag{3.59}$$

where f = 1/T is the fundamental frequency. For a sinusoidal PWM, the quantity N_s is a constant, whereas for a hysteresis band PWM it can be easily counted.

Turn-off Switching Loss

Turn-off and turn-on switching losses have been discussed and mathematically analyzed in detail by Mc Murray [32], for a DC chopper. The same formulation was extended here for the case of inverter design.

The selection of snubber capacitor typically takes into account a number of factors, such as the maximum dV/dt that the transistor can withstand, the allowable voltage overshoot during turn-off, and the turn-off switching losses. However, only the loss aspect was considered here. It can be shown [32] [37] that, for minimum total switching loss (snubber loss + turn-off loss), the value of snubber capacitor C_s is usually small, such that, during turn-off the collector-emitter voltage rise time t_{rv} is smaller than the collector current fall time t_f, as indicated in Fig. 3.12. Besides the assumption of a small C_s, another one is made in the following analysis [32]:

"During turn-off, the current fall is a linear time function completely determined

by the transistor characteristics, while the voltage rise is determined by the resulting action of the shunt snubber".

By applying this simplifying assumption, the transistor action during a turn-off can be expressed as

$$i_C = I_L\left(1 - \frac{t}{t_f}\right) \quad ; 0 \le t \le t_f \tag{3.60}$$

where I_L is the load current, i_C is the transistor collector current, and t_f the current fall time, as indicated in Fig. 3.12. Similarly, the transistor collector-emitter voltage v_{CE} is governed by the snubber action as

$$v_{CE} = \frac{1}{C_s}\int_o^t (I_L - i_c)\, d_t \tag{3.61}$$

Substituting (3.60) into (3.61) and solving the integral leads to

$$v_{CE} = \frac{I_L t^2}{2C_s t_f} \quad ; 0 \le t < t_{rv} \tag{3.62}$$

The v_{CE} rise time t_{rv} can be obtained from (3.62) by applying the end condition $v_{CE} = V_d$:

$$t_{rv} = \sqrt{\frac{2C_s V_d t_f}{I_L}} \tag{3.63}$$

Neglecting the small voltage ringing, the transistor power dissipation can be expressed as:

$$p_{t1} = \frac{I_L^2}{2C_s}\left(1 - \frac{T}{t_f}\right)\frac{t^2}{t_f} \quad ; 0 \le t < t_{rv} \tag{3.64}$$

$$p_{t2} = V_d I_L\left(1 - \frac{t}{t_f}\right) \quad ; t_{rv} \le t < t_f \tag{3.65}$$

Correspondingly, the transistor energy loss in one turn-off is

$$W_t = \int_o^{t_{rv}} p_{t1}\,dt + \int_{t_{rv}}^{t_f} p_{t2}\,dt \qquad (3.66)$$

Equation 3.63 to 3.66 indicate that the transistor turn-off loss decreases as the snubber capacitance is increased, at the expenses of augmented snubber loss. For an inverter, the load current I_L is nearly sinusoidal, and a precise computation of switching losses would require the solution of the preceding equations at every switching. By using the absolute value of the half-cycle average I_{av} of the load current in place of I_L, a good estimate of switching loss can be produced, with significant reduction in computation time. As a result, the average transistor turn-off power loss for all three phase legs P_{toff} can be expressed as

$$P_{toff} = 3\,K_{off}\,V_d\,(N_s/2)\,I_{av}\,f \qquad (3.67)$$

where $N_s/2$ is the number of lossy turn-offs per converter leg in one cycle of fundamental frequency f, and the turn-off parameter K_{off} is given by

$$K_{off} = \frac{t_f}{2}\left(1 - \frac{4}{3}\frac{t_{rv}}{t_f} + \frac{1}{2}\left(\frac{t_{rv}}{t_f}\right)^2\right) \qquad (3.68)$$

From (3.67) it can be seen that the transistor turn-off loss can be represented by an equivalent DC link shunt resistance as

$$R_{toff} = \frac{V_d}{3K_{off}(N_s/2)I_{av}f} \qquad (3.69)$$

Turn-on Switching Loss

For a transistor inverter, no snubber inductance L as indicated in Fig. 3.10 is normally used. The stray inductance due to the wiring between the DC link capacitor and the transistor acts as a parasitic inductance for turn-on snubber. The operation of the parasitic series snubber is dual to that of the shunt snubber, and by using a procedure similar to that used for turn-off loss computation, the transistor turn-on power loss can be given as [32]

$$P_{ton} = 3 K_{on} V_d (N_s/2) I_{av} f = \frac{V_d^2}{R_{ton}} \tag{3.70}$$

where the turn-on constant K_{on} is defined as

$$K_{on} = \frac{t_{fv}}{2}\left(1 - \frac{4}{3}\frac{t_r}{t_{fv}} + \frac{1}{2}\left(\frac{t_r}{t_{fv}}\right)^2\right) \tag{3.71}$$

where t_r is the collector current rise time and $t_{fv} = t_r + t_{rr}$ is the v_{CE} voltage fall time. Again, t_r is given by

$$t_r = \sqrt{\frac{2 L t_{fv} I_{av}}{V_d}} \tag{3.72}$$

From eqn. 3.70, the transistor turn-on loss can be represented by an equivalent DC link shunt resistance as

$$R_{ton} = \frac{V_d}{3 K_{on} (N_s/2) I_{av} f} \tag{3.73}$$

Equations 3.70 to 3.72 indicate that the turn-on loss decreases as the inductance L is increased.

3.4 Model Validation

Both the converter and machine models, as discussed before, were simulated in detail, using PC-SIMNON, for a 10 HP drive with indirect vector control, with all the routines included in the Appendix. The diode rectifier routine is a direct translation of the mathematical model of Section 3.3.1. The inverter model uses hysteresis-band current control (HBPWM), and some implementation details will be now discussed. For phase leg A, a logic state variable A is defined such that, if A=1 the upper transistor (Q_1) conducts, whereas if A=0, the lower transistor is turned on. The new state of the variable (NA) is defined as

$$NA = \text{If } I_{ac} > HB \text{ Then } 1 \text{ Else If } I_{ac} < -HB \text{ Then } 0 \text{ Else } A$$

where I_{ac} is the current error ($= I_a^* - I_a$) and HB is the amplitude of the hysteresis band. The phase A voltage with respect to "ground" (v_{ag}) is obtained from (3.54). The actual phase to neutral voltage (v_{an}), is given by [17]:

$$v_{an} = \frac{2}{3} v_{ag} - \frac{1}{3} v_{bg} - \frac{1}{3} v_{cg} \qquad \textbf{(3.74)}$$

The number of switchings per cycle N_s is computed by counting the transitions in the logic state variable A. The carrier frequency f_c is directly obtained by the product of N_s and the fundamental frequency f. These variables are then used in the computation of converter and machine frequency dependent parameters. Converter conduction and switching loss computation closely follows the mathematical formulation presented in

Section 3.3.

The equivalent circuits of Fig. 3.7 were used in the derivation of machine state equations. For convenience, the Q^e -axis circuit is reproduced in Fig. 3.13, indicating the selected state variables. For each axis, five state variables were selected, in this case the currents through the inductances L_{ls}, L_{ssn}', L_m, L_{lr} and L_{rsn}'. The current through L_{lm}' (i_{qrm}) does not constitute an independent state variable, since the following nodal equation must hold:

$$i_{qsl} = -i_{qsr} + i_{qrm} + i_{qm} - i_{qrl} - i_{qrr} \tag{3.75}$$

The above equation can also be written in terms of current derivatives:

$$\frac{d}{dt} i_{qsl} = -\frac{d}{dt} i_{qsr} + \frac{d}{dt} i_{qrm} + \frac{d}{dt} i_{qm} - \frac{d}{dt} i_{qrl} - \frac{d}{dt} i_{qrr} \tag{3.76}$$

In order to obtain a valid SIMNON model, the system must be described in a state equation format. To avoid unnecessary coupling among the state variables derivatives, the loop equations were carefully selected, as shown in Fig. 3.13, such that the inductance L_{ls} is present in every loop. For example, the loop through L_{ssn}' yields:

$$\frac{d}{dt} i_{qsr} = \frac{1}{L_{ssn}'} \left[L_{ls} \frac{d}{dt} i_{qsl} - R_{ssn}' i_{qsr} \right] \tag{3.77}$$

Substitution of (3.77) along with the other loop equations into (3.76) produced the state equation for i_{qsl}. The remaining state equations were readily obtained by further substitution of the right side of di_{qsl}/dt into the corresponding loop equations.

Both steady-state and transient conditions of the system were considered in the validation process of the models. Table 3.1 shows the complete power circuit parameters of the drive. The steady-state system performance was initially investigated, for various

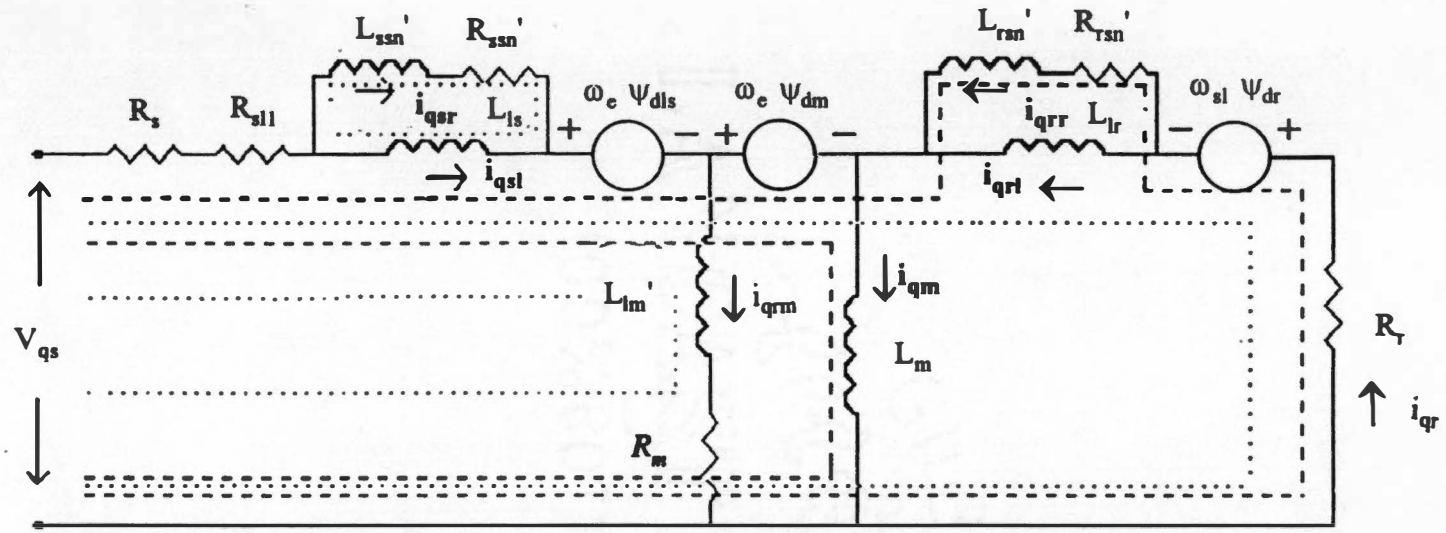


Fig. 3.13 Q^e-axis circuit showing loops for state equations derivation.

Table 3.1 Power circuit parameters of the AC drive system.

Machine: 10 hp 230 / 460 V 27/13.5 A
1755 rpm 60 Hz Class B 1.15 SF

D^e-Q^e equivalent circuit parameters

R_s=0.2264 Ω	R_r=0.1256 Ω	R_m=129.06 Ω
L_{ls}=0.00155 H	L_{lr}=0.00193 H	L_m=0.0275 H
R_{sll}=0.0341 Ω	R_{ssn}'=146.52 Ω	R_{rsn}'=169.63 Ω
L_{lm}'=0.0032 H	L_{ssn}=7.75 10^{-5} H	L_{rsn}=9.65 10^{-5} H

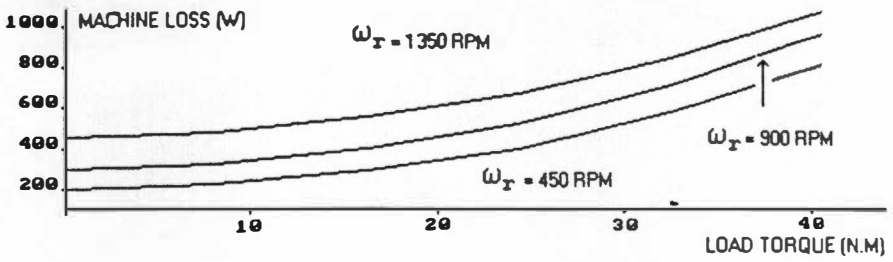
Diode: POWEREX CD 411230 Dual diode module
30 A / 1200 V
Drop equation parameters:
v_{d0}=0.8 V K=0.052 m=0.585

Transistor: POWEREX KS 524503 Single Darlington transistor module, 30 A / 600 V
Drop equation parameters:
v_{t0}=0.7 V R_t=0.020

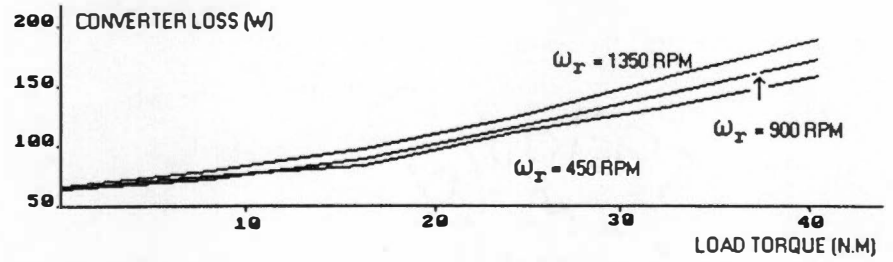
load torque and speed conditions, and the results are shown in Fig.3.14. From the machine loss profiles of Fig. 3.14(a) it can be seen that, for given speed, the total loss increases with torque, primarily due to increased fundamental copper and stray load losses. For a constant load torque, the losses increase with speed, mainly because of additional core loss and friction and windage losses. Fig. 3.14(b) shows the corresponding total converter loss for the same load torque and speed conditions. It can be seen that the converter loss is more affected by an increase in load torque at constant speed, rather than by increase of machine speed at constant load torque. This can be explained as follows:

With rated flux, the machine current is essentially a function of load torque and is practically independent of speed. Therefore, the inverter losses, that basically depend on machine current, is dominantly influenced by load torque. On the other hand, the diode rectifier loss is a function of DC link current, that increases with converter output power. Therefore, the rectifier loss is influenced by both speed and torque of the machine. Fig. 3.14(c) shows the total system efficiency at various torque and speed conditions. It can be noticed that the point of maximum efficiency moves toward the rated torque, as the speed increases, since the machine is designed to yield maximum efficiency at rated torque and speed. The upper speed used in the study is less than the base value of 1800 rpm, because the DC link voltage is not sufficiently high to enforce pure PWM operation near base speed, what is required for correct operation of the vector control system.

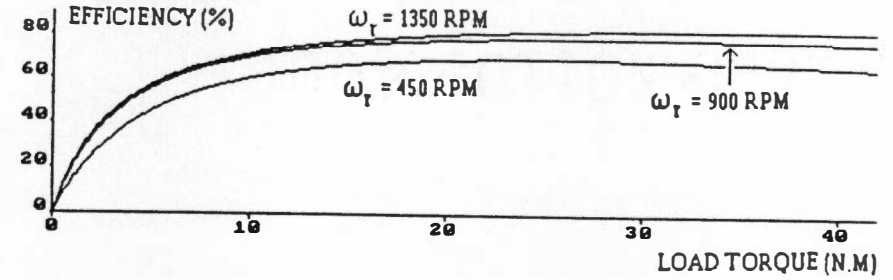
The transient performance will now be discussed. Fig. 3.15(b) and (c) show the flux responses to the magnetizing command current i_{ds}^* step of Fig. 3.15(a). The response



(a)



(b)



(c)

Fig. 3.14 Steady state performance.
a) Machine loss at various torques and speeds.
b) Converter loss at various torques and speeds.
c) System efficiency at various torque and speed.

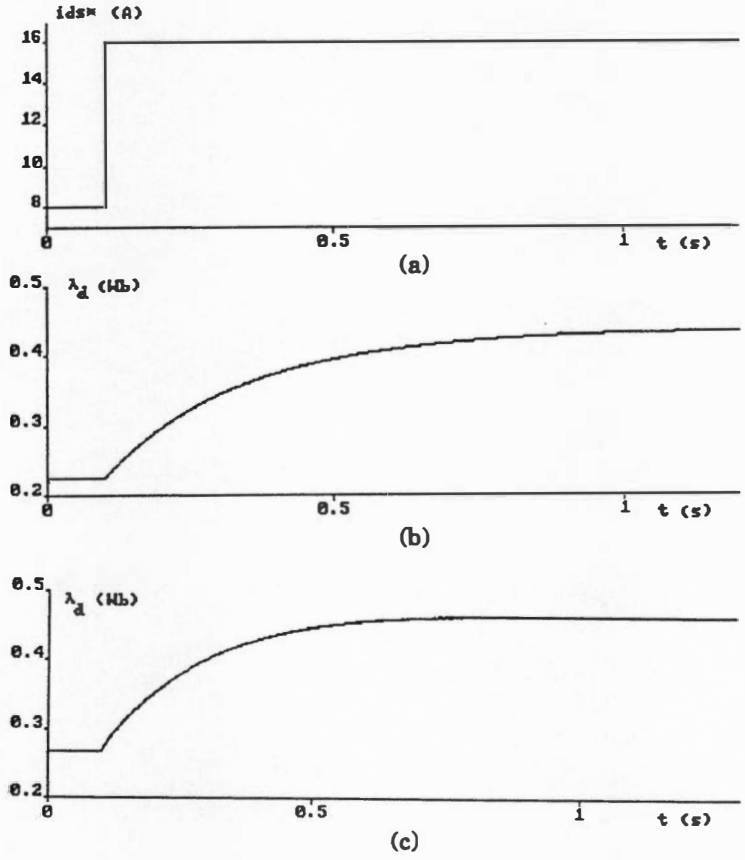


Fig. 3.15 Rotor flux responses at constant speed.
a) Command current step.
b) Standard D^e -Q^e model response.
c) Lossy D^e -Q^e model response.

of the standard D^e - Q^e model of Fig. 3.2 is initially shown in Fig. 3.15(b), whereas that of the lossy model of Fig. 3.7 is presented in Fig. 3.15(c). It can be seen that the response of the proposed model has a smaller rise time than that of the standard model. This is caused by the following factors:

- There is an increase in rotor resistance due to temperature effect, included in the proposed model, but neglected in the standard model.
- The shunt branches that represent core loss and rotor stray loss have a transient effect of increasing the effective rotor resistance while decreasing the equivalent rotor leakage inductance. Both factors combine to reduce the effective rotor time constant ($\tau_r = L_r/R_r$). While the amount of reduction in τ_r is dependent on the proper estimation of machine parameters, it is consistent from a theoretical point of view.

Another aspect that can be observed is that, the steady-state value of rotor flux is somewhat smaller for the proposed model. This reflects the saturation effect on the magnetizing inductance, included in the proposed model, but ignored in the standard model.

Finally, the transient torque response of an indirect vector-controlled drive was investigated at constant speed, as indicated in Fig. 3.16, to study possible effect for D^e-Q^e equivalent circuits modification. Fig. 3.16(a) shows the step in the torque component of current (i_{qs}^*) at the rated flux condition. Fig. 3.16(b) shows the corresponding torque response for lossless converter and ideal D^e-Q^e machine model with slip gain parameter tuned with the nominal machine parameters. The observed rise time is essentially due to

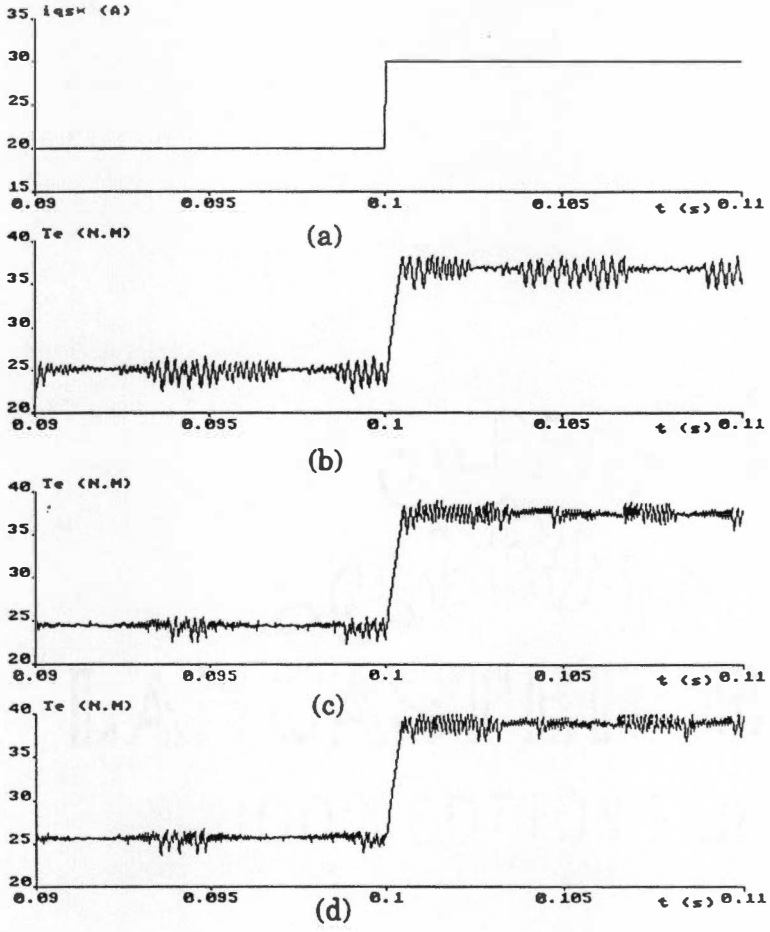


Fig. 3.16 Torque responses of the drive at rated flux (speed = 900 RPM).
a) Command current step.
b) Ideal and lossless converter-machine model (slip gain parameters are nominal machine parameters).
c) Lossy converter-machine model (slip gain parameters are nominal machine parameters).
d) Lossy converter-machine model (slip gain parameters track with machine parameters).

intrinsic delay of hysteresis-band current controller. Fig. 3.16(c) shows the response for lossy converter machine model, with nominal machine parameters used in the slip gain, whereas in Fig. 3.16(d) the same model was used, except for the slip gain, that is tuned for the actual machine parameters at the particular operating condition. The responses for all the three conditions are practically identical, what reflects a small mismatch between actual and nominal machine parameters used in the slip gain computation, for this particular operating condition.

In conclusion, the loss modeling of converter induction machine system proposed here takes into consideration the relevant losses occurring in a practical drive system. The simulation study has demonstrated the coherent system behavior at both steady-state and dynamic conditions. The actual accuracy of the models is dependent on the precision with which the machine and converter parameters can be obtained. However, the topic of parameter measurement or estimation is beyond the scope of this work.

CHAPTER 4

FUZZY EFFICIENCY OPTIMIZATION CONTROL

4.1 Introduction

Efficiency improvement in variable frequency drives is getting a lot of attention in the recent years. Higher efficiency is important not only from the viewpoints of energy saving and its obvious financial payoff, but also from the broad perspective of environmental pollution control. In fact, as the use of variable speed drives continues to increase in areas traditionally dominated by constant speed drives, motivated by the quest for increased productivity, the financial and environmental payoffs reach new importance.

The efficiency of a drive system is a complex function of the type of machine used, converter topology, type of power semiconductor switches and the selected PWM algorithm. In addition, the control system has profound effect on the drive efficiency. A drive system normally operating at rated flux gives the best transient response. However, at light loads, rated flux operation causes excessive core loss, compared to copper loss, thus impairing the efficiency of the drive. Since drives operate at light load most of the time, optimum efficiency can be obtained by flux control.

A number of methods for efficiency improvement through flux control have been proposed in the literature, since its principle was first introduced by Nola [38]. They can be classified into three basic types:

1) The simple pre-computed flux program as a function of torque, that is widely used for light load efficiency improvement. The scheme can be improved by generating the flux program at discrete speeds, to take the frequency dependency into consideration. This method, however, yields only a partial improvement in the system efficiency.

2) The real time computation of losses and corresponding selection of flux level that results in minimum losses is quite elegant. However, as the loss computation is based on a machine model, parameter variations caused by temperature and saturation effects tend to yield sub-optimal efficiency operation [60].

3) The on-line efficiency optimization control [39]-[41] on the basis of search, in which the flux is reduced in steps until the measured input power settles down to the lowest value. The control does not require the knowledge of machine parameters, is completely insensitive to parameter changes, and the algorithm is applicable universally to any arbitrary machine.

Induction motors are by far the most used among all electric motors, and are responsible for the consumption of a large fraction of the total electric energy produced. It is, therefore, natural that they have been the primary focus of the efficiency improvement studies. Part of this effort has been dedicated to vector controlled induction motor drives, since they possess inherent decoupling of flux and torque control loops, making them well suited for efficiency optimization control. In fact, Kirschen [40] investigated such system, using on line search method. Fig. 4.1 shows a typical power minimization process for a partial load condition. As the command flux is decreased, the

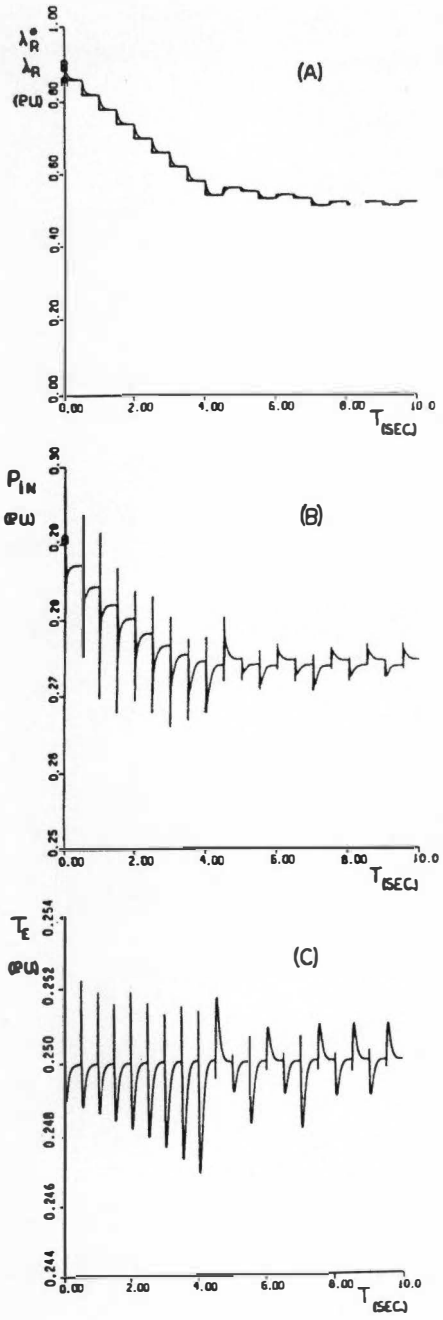


Fig. 4.1 Simulation of the input power minimization process [40].

input power initially decreases and then settles to the minimum value. There is, however, an undesirable side effect: as the flux is reduced, the developed torque is also affected, as demonstrated by the torque pulses of Fig. 4.1(c). If the load inertia is small, the low frequency torque pulsation will translate in speed fluctuation, that can eventually lead to mechanical resonance.

In this chapter, a fuzzy logic based on-line efficiency optimization control for an indirect vector controlled drive system is discussed. Fast convergence has been achieved by using adaptive step size of the excitation current. The low frequency pulsating torque generated by the efficiency controller has been suppressed by a feed-forward compensation algorithm. Both simulation and experimental studies were performed to validate the theoretical development. A paper on this subject [62] has been accept for presentation at the 1993 international conference on industrial electronics, control, and instrumentation, IECON' 93.

4.2 Fuzzy Efficiency Optimization of a Vector Control Drive

An indirect vector controlled induction motor drive incorporating the proposed efficiency optimization controller is shown in Fig. 4.2. It consists of a diode rectifier and a hysteresis band PWM transistor inverter. All the control functions indicated by the dashed outline are implemented in real time by a single digital signal processor (TMS320C25 from Texas Instrument, Inc.). The feedback speed control loop generates the active or torque current command $i_{qs}^{*'}$, as indicated. The vector rotator receives the torque and excitation current commands i_{qs}^{*} and i_{ds}^{*}, respectively, from the two positions

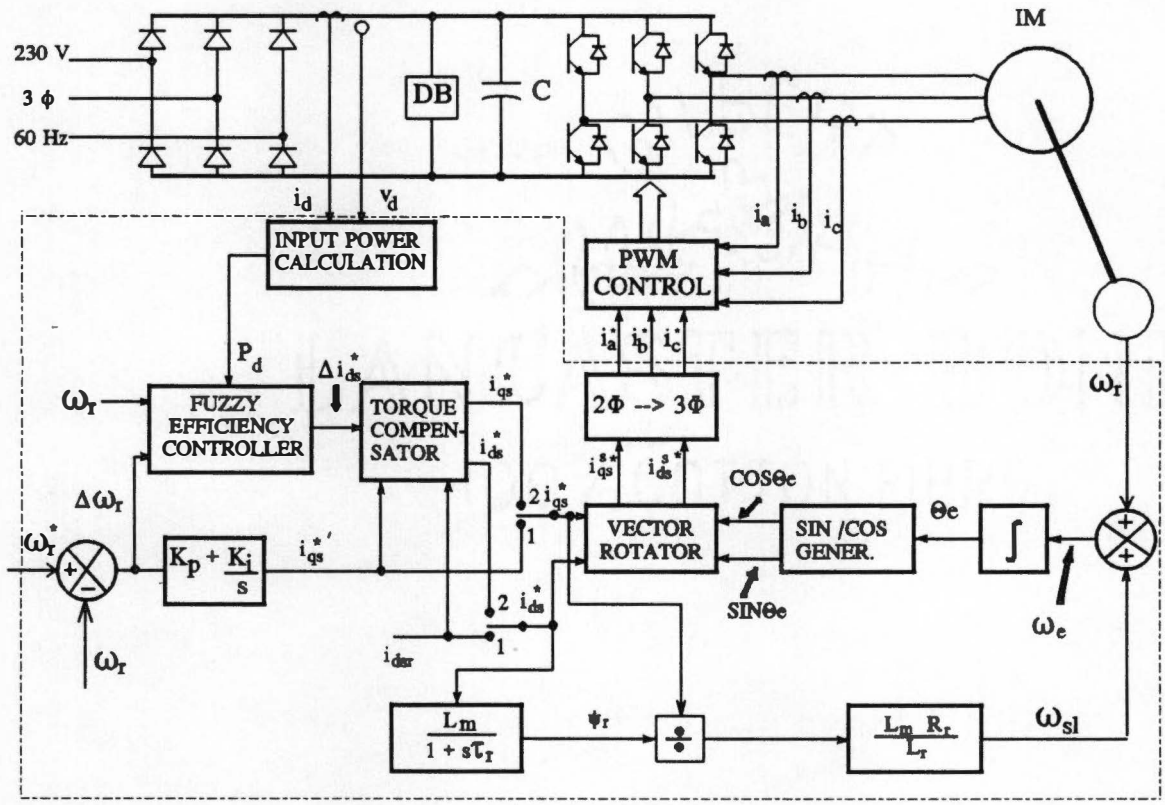


Fig. 4.2 Indirect vector controlled induction motor drive incorporating the efficiency optimization controller.

of a switch: the transient position (1), where the excitation current is established to the rated value i_{dsr} and the speed loop feeds the torque current; and the steady-state position (2), where the excitation and torque currents are generated by the fuzzy efficiency controller and feedforward torque compensator which will be explained later. The fuzzy controller becomes effective at steady-state condition, i.e., when the speed loop error $\Delta\omega_r$ approaches zero. Note that the DC link power P_d, instead of input power, has been considered for the fuzzy controller since both follow symmetrical profiles. This can be demonstrated as follows:

- Minimization of P_d translates into minimization of DC link current i_d, since the DC link voltage v_d is essentially constant;
- Rectifier loss are proportional to i_d , (Eqn. 3.50);
- Minimum i_d yields minimum rectifier loss, and consequently, minimum input power.

The principle of efficiency optimization control with rotor flux programming at a steady-state torque and speed condition is explained in Fig. 4.3. The rotor flux is decreased by reducing the magnetizing current, which ultimately results in a corresponding increase in the torque current, normally by action of the speed controller, such that the developed steady-state torque remains constant. As the flux is reduced, the iron loss decreases with the attendant increase of copper loss. However, the total system loss (converter and machine) decreases, resulting in a reduction of DC link power. The search is continued until the system settles down at the minimum input power point A, as indicated. Any excursion beyond the point A will force the controller to return to the

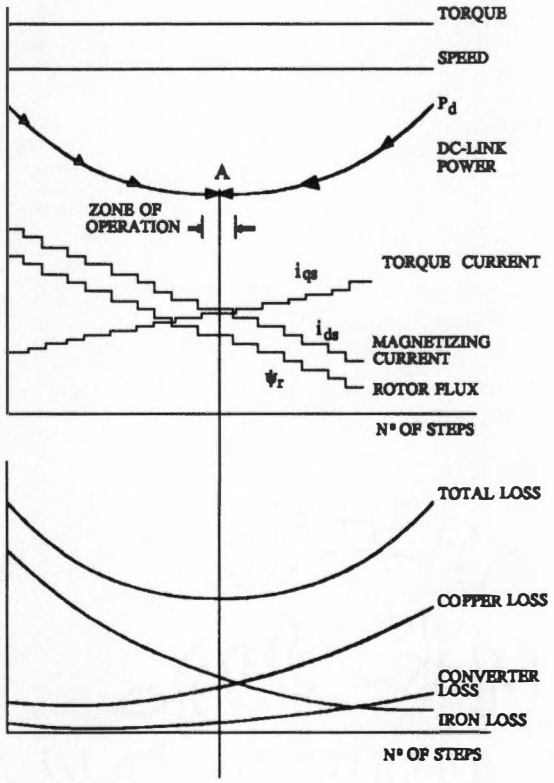


Fig. 4.3 Principle of efficiency optimization control with rotor flux programming.

minimum power point.

4.2.1 The Efficiency Optimization Controller

The fuzzy efficiency controller operation is explained in Fig. 4.4. The DC link power is sampled and compared with the previous value to determine the increment ΔP_d. In addition, the last excitation current decrement $L\Delta i_{ds}$ is reviewed. On these basis, the decrement step of Δi_{ds}^* is generated from fuzzy rules through fuzzy inference and defuzzification [14], as indicated. The adjustable gains P_b and I_b, generated by the scaling factors computation block, convert the input variable and control variable, respectively, to per unit values, such that a single fuzzy rule base can be used for any torque and speed condition. The input gain P_b as a function of machine speed ω_r can be given as

$$P_b = a\omega_r + b \tag{4.1}$$

The output gain I_b is computed from the machine speed and an approximate estimate of machine torque $\hat{T}_e$ as

$$I_b = c_1\omega_r - c_2\hat{T}_e + c_3 \tag{4.2}$$

where

$$\hat{T}_e = K_t' i_{ds}^* i_{qs}^* \tag{4.3}$$

The appropriate coefficients a, b, c_1, c_2, and c_3 were derived from simulation studies as follows:

. Using simplified models, the approximated optimum magnetizing current i_{dso} was determined for selected operating points in the torque-speed plane;

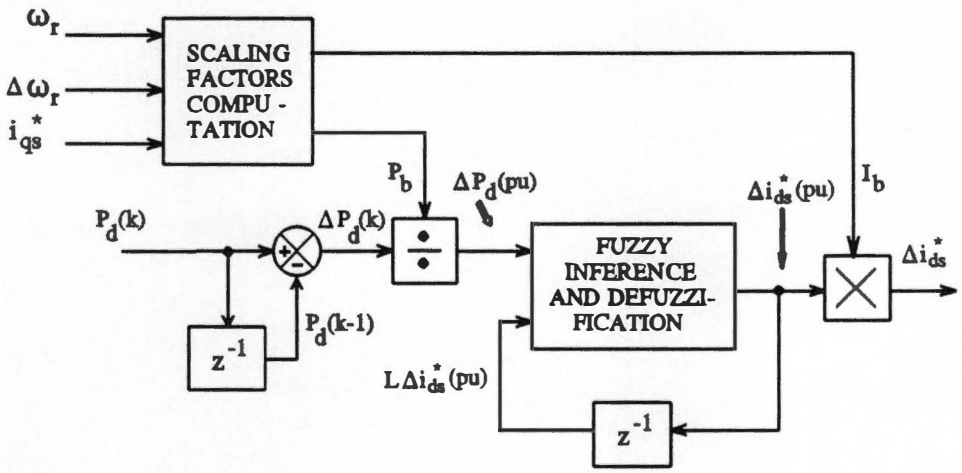


Fig. 4.4 Efficiency optimization control block diagram.

- Least-square estimation was used to obtain a linear representation of i_{dso} as a function of machine torque T_e and speed ω_r;
- The output gain expression (4.2) was defined as one third of the distance between rated magnetizing current i_{dsr} and approximate optimum i_{dso}.
- A new set of simulations was performed, for a load torque of 0.1 pu and different speeds. The magnetizing current was reduced by a step size I_b given by (4.2), and the corresponding reduction in input power ΔP was obtained.
- The input power expression (4.1) was defined as the linear representation of ΔP^{-1} as a function of speed, derived by least square estimation.

A few words on the importance of the input and output gains are appropriate here. In the absence of input and output gains, the efficiency optimization controller would react equally to a specific value of ΔP_d, resulting from a past action $\Delta i_{ds}^*(k-1)$, irrespective of operating speed. Since the optimal efficiency point A (Fig. 4.3) is speed dependant, the control action could easily be either too conservative, resulting in slow convergence, or excessive, yielding an overshoot in the search process, with possible adverse impact on system stability. As both input and output gains are functions of speed, this problem does not arise. Eqn. 4.2 also incorporates the "a priori" knowledge that the optimum value of i_{ds}^* is a function of torque as well as machine speed. In this way, for different speed and torque conditions, the same $\Delta i_{ds}^*(pu)$ will result in different Δi_{ds}^*, ensuring a fast convergence. One additional advantage of per unit basis operation is that the same fuzzy controller can be applied to any arbitrary machine, by simply changing the coefficients of input and output gains.

The membership functions for the fuzzy efficiency controller are shown in Fig. 4.5. Due to the use of input and output gains, the universe of discourse for all variables are normalized in the [-1,1] interval. It was verified that, while the control variable Δi_{ds}^* required 7 fuzzy sets to provide good control sensitivity, the past control action $L\Delta i_{ds}^*$, i.e., $\Delta i_{ds}^*(k-1)$, needed only 2 fuzzy sets, since the main information conveyed is the sign. The small overlap of the positive (P) and negative (N) membership functions is required to ensure proper operation of the height defuzzification method, i.e., to prevent indeterminate result in case $L\Delta i_{ds}^*$ approaches zero.

The rule base for fuzzy control is given in Table 4.1. An example fuzzy rule can be given as:

IF the power increment (ΔP_d) is negative medium (NM) and the last Δi_{ds}^* ($L\Delta i_{ds}$) is negative (N), THEN the new excitation increment (Δi_{ds}^*) is negative medium (NM).

The basic idea is that, if the last control action indicated a decrease of DC link power, proceed searching in the same direction, and the control magnitude should be somewhat proportional to the measured DC link power change. In case the last control action resulted in an increase of P_d ($\Delta P_d > 0$), the search direction is reversed, and the Δi_{ds}^* step size is reduced to attenuate oscillations in the search process.

As the optimum point varies with speed and load conditions, and is also affected by changes in machine parameters, the fuzzy logic search controller constitutes a natural choice for this problem. Initially, a linguistic description of the control strategy could be easily obtained. Fuzzy logic then provided the appropriate mathematical framework for the derivation of the actual controller. The fine tuning process was, nevertheless,

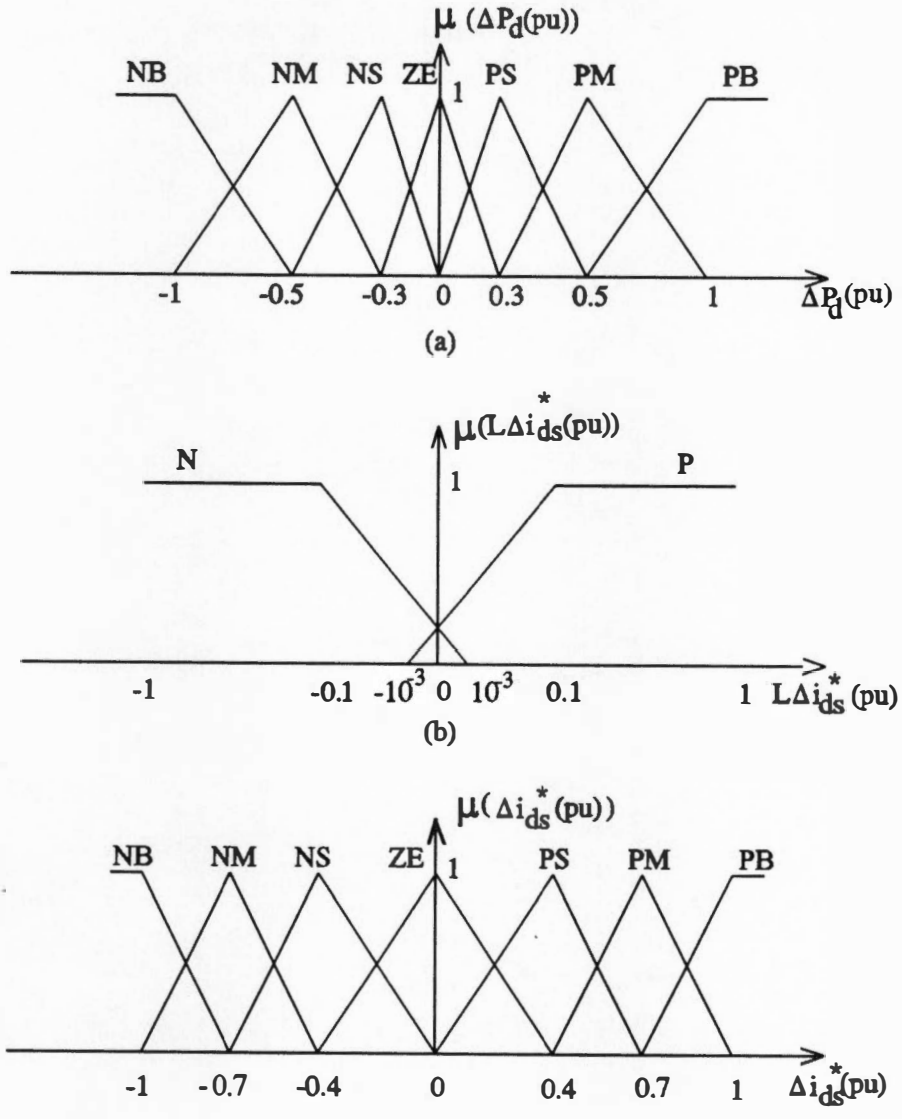


Fig. 4.5 Membership functions for the fuzzy efficiency control.
(a) Change in DC link power (ΔP_d(pu)).
(b) Last change in magnetizing current ($L\Delta i_{ds}^*$(pu)).
(c) Magnetizing current increment (Δi_{ds}^*(pu)).

Table 4.1 Rule base for the fuzzy efficiency controller.

$L\Delta i_{ds}(pu)$ / $\Delta P_d(pu)$	N	P
PB	PM	NM
PM	PS	NS
PS	PS	NS
ZE	ZE	ZE
NS	NS	PS
NM	NM	PM
NB	NB	PB

somewhat time consuming.

4.2.2 Feed-forward Pulsating Torque Compensation

As the excitation current is reduced in adaptive steps by the fuzzy controller, the rotor flux ψ_r decreases exponentially [17], which is given by

$$\frac{d}{dt}\psi_{dr} = \frac{L_m i_{ds} - \psi_r}{\tau_r} \tag{4.4}$$

where τ_r (= L_r/R_r) is the rotor time constant and L_m the magnetizing inductance. The decrease of flux causes loss of torque, which normally is compensated slowly by the speed control loop. Such pulsating torque at low frequency is very undesirable because it causes speed ripple and may create mechanical resonance. To prevent these problems, a feed-forward pulsating torque compensator is proposed here.

Under correct field orientation control, the developed torque is given by

$$T_e = K_t i_{qs} \psi_r \tag{4.5}$$

For an invariant torque, the torque current i_{qs} should be controlled to vary inversely with the rotor flux ψ_r. A practical implementation of this concept is illustrated in Fig. 4.6. A compensating signal $\Delta i_{qs}^*(t)$ is added to the original $i_{qs}^{*'}$ to counteract the decrease in flux $\Delta\psi_r(t)$, where $t \in [0,T]$ and T is the sampling period for efficiency optimization control. Let $i_{qs}(0)$ and $\psi_{dr}(0)$ be the initial values for i_{qs} and ψ_d, respectively, for the k-th step change of i_{ds}^* (k=2 in Fig. 4.6.) For a perfect compensation, the developed torque must

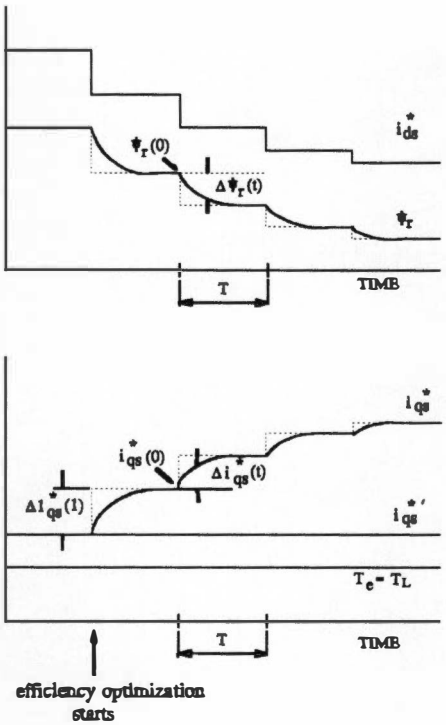


Fig. 4.6 Principles of feedforward torque compensation.

remain constant, and the following equality holds

$$[\psi_r(0) + \Delta\psi_r(t)][i_{qs}(0) + \Delta i_{qs}(t)] = \psi_r(0)\, i_{qs}(0) \tag{4.6}$$

Solving for $\Delta i_{qs}(t)$ yields

$$\Delta i_{qs}(t) = \frac{-\Delta\psi_r(t)\, i_{qs}(0)}{\psi_r(0) + \Delta\psi_r(t)} \tag{4.7}$$

where $\Delta\psi_r(t)$ is governed by eqn. 4.4, with Δi_{ds} substituted for i_{ds}. To implement such compensation, eqn. 4.7 is adapted to produce $\Delta i_{qs}^*(t)$, using flux estimate $\hat{\psi}_r$ and command i_{qs}^* in place of actual signals. A good approximate solution for $\Delta i_{qs}^*(t)$ can be obtained by replacing the denominator of (4.7) by its estimated steady-state value $\hat{\psi}_r(T)$ [$= L_m\, i_{ds}^*(T)$]. In this case, the compensation can be implemented in two steps as shown in Fig. 4.7. First, the value for the compensating current step is computed by discrete eqn. 4.8.

$$\Delta I_{qs}^*(k) = \frac{\hat{\psi}_r(k-1) - \hat{\psi}_r(k)}{\hat{\psi}_r(k)}\, i_{qs}^*(k-1) \tag{4.8}$$

Next, the current step is processed through a first order low pass filter of rotor time constant, and then added to the previous compensating steps, as also illustrated in Fig. 4.6. This current is added to the original speed loop generated current $i_{qs}^{*'}$, so that, at any instant the product $i_{qs}\, \psi_r$ remains essentially constant. It may be noted, however, that the actual rotor time constant of the machine may vary, giving somewhat imperfect pulsating torque compensation.

To ensure adequate compensation, magnetizing inductance saturation effects were taken into account using a piece-wise linear model, similar to that developed for the induction machine model. The magnetizing inductance L_m is estimated on line from the

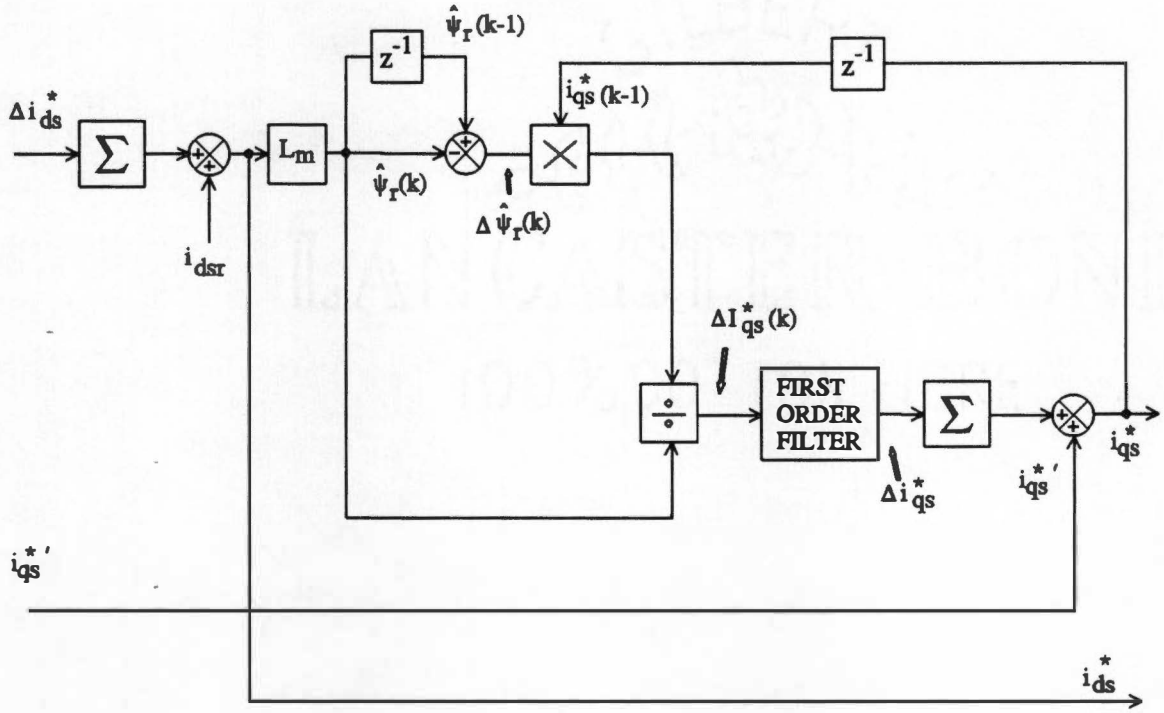


Fig. 4.7 Feedforward pulsating torque compensator block diagram.

command i_{ds}^* as

$$\begin{aligned} L_m &= L_{m0} && ,if\ i_{ds}^* \le i_{ds0} \\ L_m &= L_{m0} - m_d (i_{ds}^* - i_{ds0}) && ,if\ i_{ds}^* > i_{ds0} \end{aligned} \tag{4.9}$$

where L_{m0} is the non-saturated value of L_m, i_{ds0} is the linear region breakpoint, and m_d is a saturation coefficient. All parameters are obtainable from a standard no-load test.

4.2.3 Transition to Optimum Transient Response Mode

It may me noted from the preceding discussion that efficiency optimization control is only effective at steady-state condition. A disadvantage of this control mode is that the transient response becomes sluggish, because the weaker flux reduces the maximum torque capability, as indicated by eqn. 4.5. For any change in load torque or speed command, fast transient response capability of the drive can be restored by establishing the rated flux. Fig. 4.8 shows the criteria for transition between efficiency optimization (Mode 2) and transient response optimization (Mode 1). The system starts in Mode 1 and then switches to Mode 2 when the steady-state criteria is met. In the event of a load disturbance or a change in set speed, the system switches back to Mode 1 by establishing the rated magnetizing current.

At steady-state, current must be kept within the thermal limits of the machine and inverter, that is usually the rated machine current. During a transition from Mode 2 to Mode 1, this limit can be temporarily increased, (typically by 50% for 60 secs.) such that more torque is available to meet the acceleration / deceleration demand. This constraint is given by the relation

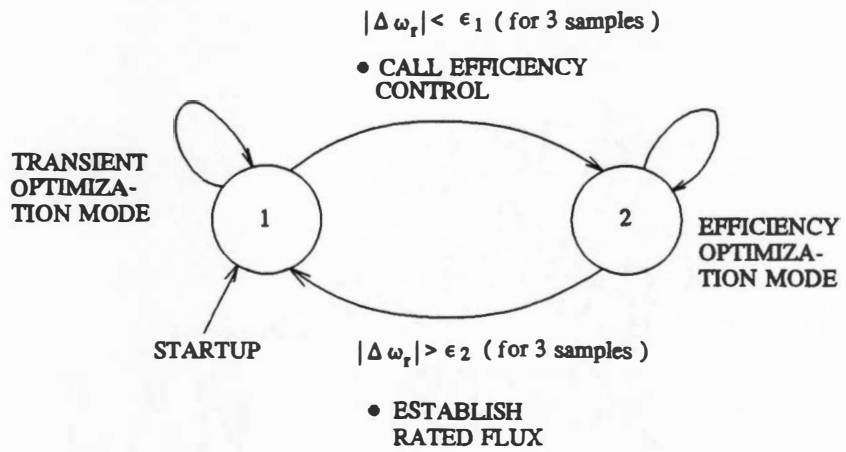


Fig. 4.8 Transition between efficiency optimization and transient response optimization modes.

$$\sqrt{i_{ds}^{*2} + i_{qs}^{*2}} < I_{lim} \tag{4.10}$$

where $I_{lim} = 1.5\ I_{sr}$, and I_{sr} is the rated stator current.

Notice that, as the rotor flux does not change instantly, the maximum torque capability following a transition might not be enough to overcome the torque demand, even with the relaxed current limit. In such scenario, there is a possibility that the drive will stall, and preventive measures should be taken. One simple solution consists in imposing a minimum value for i_{ds}^{*}, during Mode 2, that would keep the maximum torque capability at a safe level, at the expenses of suboptimum efficiency operation.

4.3 Simulation Program Development

In chapter 3, SIMNON routines for converter and induction motor lossy models were developed and integrated to a speed control system, using indirect vector control. In this section the Fuzzy Efficiency Optimization Routine (FEOPT) and Feed Forward Torque Compensation Routine (FFTC) development will be discussed. The simulation block diagram of Fig. 4.9 shows the various routines used in the study, and the associated input/output variables for every subsystem. Discrete subsystems are identified by a D whereas continuous subsystems are marked with a C.

The flowchart of Fig. 4.10 shows the structure of the FEOPT routine. For convenience, the determination of the operating mode (Mode 1 or Mode 2 of Fig. 4.8) is included in the FEOPT routine. First, the program checks the previous operating mode and verifies if any transition is required. If after the tests the system is found to be in the

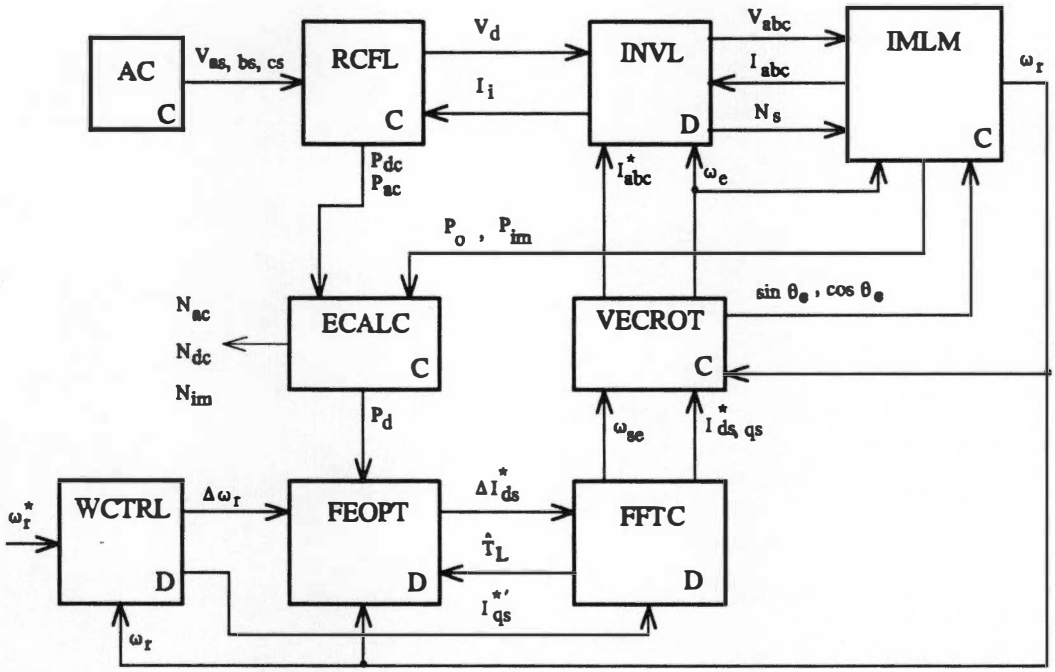


Fig. 4.9 SIMNON simulation block diagram showing I/O variables.

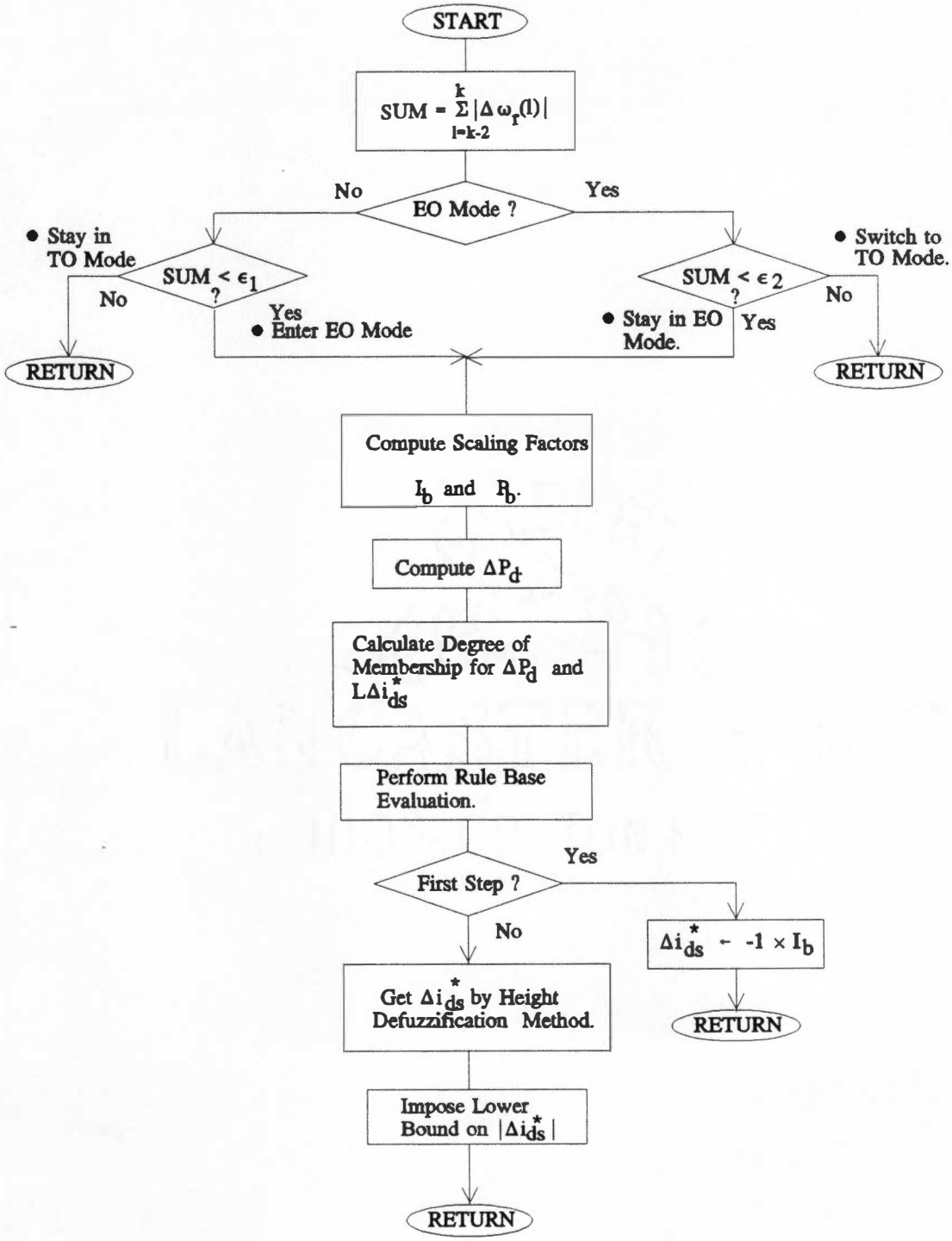


Fig. 4.10 Fuzzy efficiency optimization flowchart.

efficiency optimization (EO) mode, then the program proceeds to the tasks actually relatedto the efficiency optimization process, as indicated in the flowchart. In order to initiate the efficiency optimization process, following a Mode 1 to Mode 2 transition the first Δi_{ds}^* is set to -1 pu. The actual value in Amperes is determined by the output scaling factor I_b. From the second iteration on, Δi_{ds}^* value is truly derived from the fuzzy controller. After the optimum point is reached, a minimum (non-zero) value is imposed to Δi_{ds}^*, to keep the search process active, and ensure true optimum operation in case of parameter variation or small change in load condition.

The algorithm for fuzzy control implementation has already being described in detail in Section 2.2.1, for the case of fuzzy $\Delta\alpha$ compensation. However, its implementation using SIMNON language is somewhat complicated, mainly because SIMNON does not support branch instructions. Furthermore, the order of program execution is not completely controlled by the user, since during compilation the program sorts the instructions to optimize for execution time.

The methodology used in the actual derivation of fuzzy efficiency optimization SIMNON routine is explained here with the help of Fig. 4.11.

1. Compute input and output scaling factors, P_b and I_b, respectively, and the change in input power, ΔP_d (pu);
2. Make a preliminary computation of degree of membership for all fuzzy sets. For instance, the degree of membership of ΔP_d (k) in the fuzzy set PM, shown in Fig. 4.11 is given by:

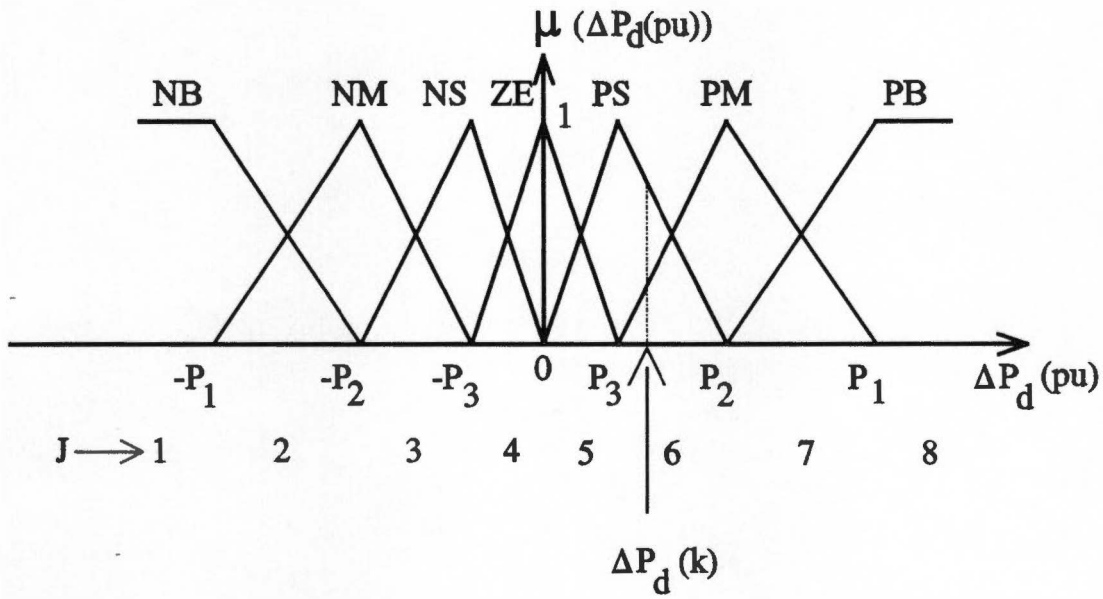


Fig. 4.11 Evaluation of degree of membership.

$$\mu_{PM}(\Delta P_d(k)) = \frac{\Delta P_d(k) - P_3}{P_2 - P_3} \qquad (4.11)$$

3. Define the interval index J that identifies which fuzzy sets possess non zero degree of membership. SIMNON supports the conditional assignment of value to a variable, but does not allow a continuation line. This problem can be solved by defining some auxiliary variables (J_1, J_2,...) as indicated below:

   J = If $\Delta P_d < -P_1$ Then 1 Else If $\Delta P_d < -P_2$ Then 2 Else J_1

   J_1 = If $\Delta P_d < -P_3$ Then 3 Else If $\Delta P_d < 0$ Then 4 Else J_2

   ...

4. Evaluate the degree of membership for the relevant fuzzy sets:

   μ_{P1} = If J < 3 Then μ_{NB} Else If J < 4 Then μ_{NM} Else... Else μ_{PB}

   $\mu_{P2} = 1 - \mu_{P1}$

   For the situation illustrated in Fig. 4.11, $\mu_{P1} = \mu_{PS}$ and $\mu_{P2} = \mu_{PM}$.

5. Apply similar procedure to compute degrees of membership μ_{I1} and μ_{I2}, for the second input variable, $L\Delta i_{ds}^*$(pu).

6. Evaluate the antecedent of the relevant rules, using MIN operator. Typically, four rules are fired. The first rule is illustrated here:

   μ_{RA} = MIN (μ_{P1}, μ_{I1})

7. Retrieve the contribution of each fired rule, from the rule base:

   ΔI_a = If J<3 Then I_1 Else If J<4 Then I_2 Else...

   Due to the symmetry of this particular rule base, $\Delta I_b = -\Delta I_a$.

8. Calculate the new Δi_{ds}^* by height defuzzification method.

$$\Delta i_{ds}^{*}(pu) = \frac{\mu_{RA}\,\Delta I_a + \ldots + \mu_{RD}\,\Delta I_d}{\mu_{RA} + \ldots + \mu_{RD}} \qquad (4.12)$$

The feed forward torque compensation (FFTC) routine operation is explained by the flowchart of Fig. 4.12. Its principle has already being discussed in section 4.2.2, and therefore, only some relevant aspects will be discussed here. During the search process (EO mode), the test for maximum stator current is based on the steady-state limit I_{liml}, that corresponds to the rated motor current. If it is detected that the new control action Δi_{ds}^{*} would result in a violation of such limit, the new magnetizing current $i_{ds}^{*}(k)$ is reset to its previous value $i_{ds}^{*}(k-1)$, and similarly, no correction is made in $i_{qs}^{*}(k)$. If the program detects that the system has switched to a TO mode (Mode 1), rated i_{ds}^{*} is reestablished and the accumulated compensating signal $\Sigma\Delta i_{qs}^{*}(k)$ is reset to zero. The actual compensating signal $\Sigma\Delta i_{qs}$ will approach zero following the first order filter dynamics. This action ensures that no sudden change in machine torque will take place, following transition in operating mode. Furthermore, as the original torque current command $i_{qs}^{*'}$ from the speed controller immediately affects i_{qs}^{*}, proper control action to any disturbance is ensured. For the TO mode, the stator current limit is set to 50% higher than its steady-state value, as mentioned before. Besides the functions related to torque compensation, the routine also computes rotor flux estimate, performs slip frequency calculation and is also responsible for approximate machine torque estimation, required by the FEOPT routine.

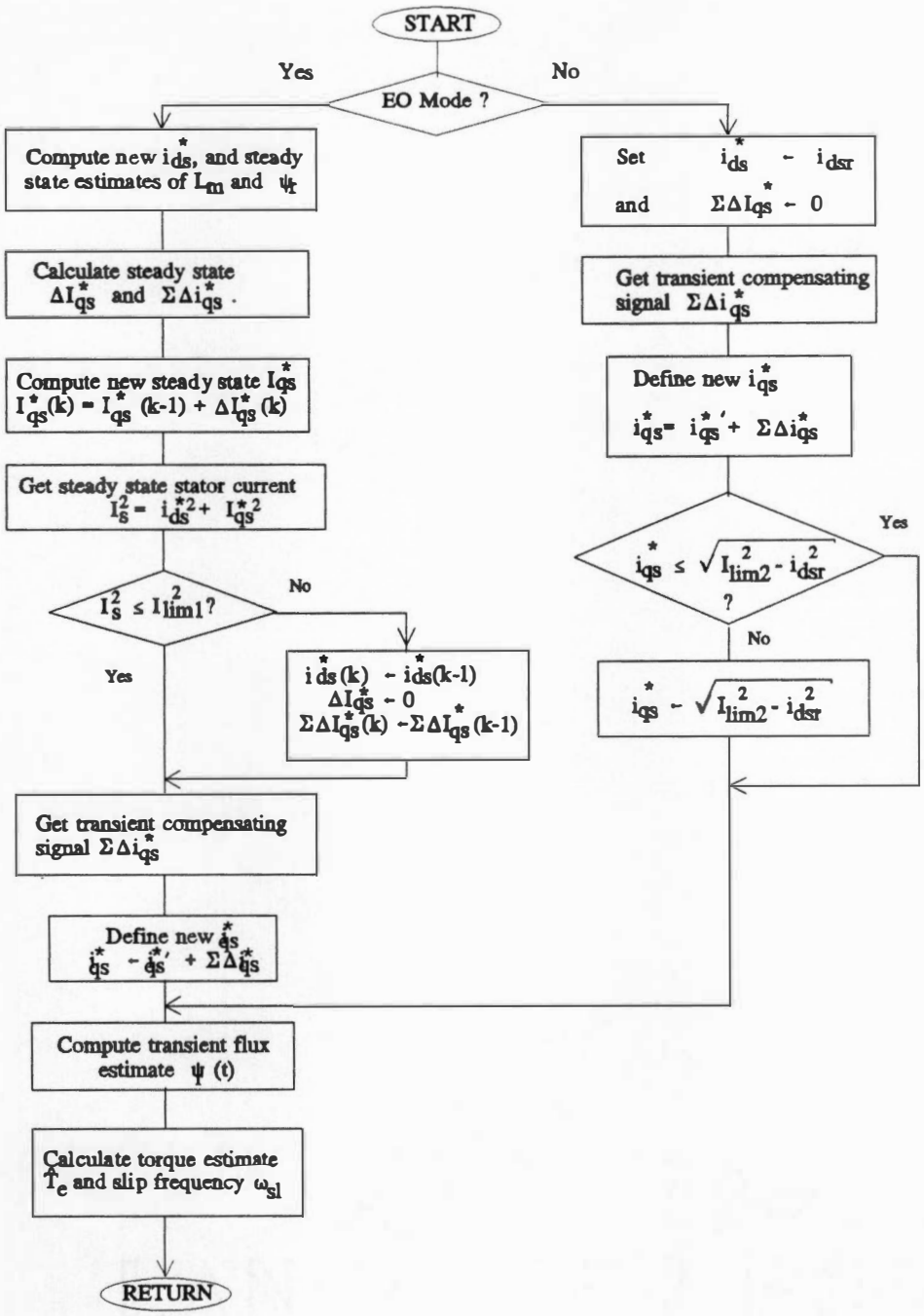


Fig. 4.12 Feedforward torque compensation simulation flowchart.

4.4 Simulations Study

After the development and tests of the SIMNON routines, a detailed performance evaluation study was conducted, using the machine parameters given in Table 4.2. Fig. 4.13 shows the time domain optimum efficiency search curves at speed ω_r = 0.25 pu and load torque T_L = 0.1 pu. The optimum flux level is achieved in only four steps of i_{ds}^*. The search algorithm, however, imposes a minimum non-zero step size of Δi_{ds}^* when ΔP_d approaches zero, in order to ensure true optimum operation for parameter variation or small change in load condition. The effectiveness of the pulsating torque compensation is evident from the torque profile of Fig. 4.13(d) and speed response of Fig. 4.13(e). Fig. 4.14 gives efficiency curves at different operating points on torque - speed plane, with and without efficiency optimization control. As the class B induction machine was designed for 230 V, 60 Hz operation, the highest speed was limited to 0.75 pu (1350 rpm), because the DC link voltage is not high enough to enforce current regulation at higher speeds, under rated flux conditions. As expected, maximum efficiency gains occur at light load, where the rated flux core losses are very dominant. Fig. 4.15 illustrates system performance during operating mode transitions. The system is initially at steady-state and the E.O. mode is activated. At t=3 sec. the system has reached the optimum efficiency point, when the load torque is suddenly increased from 0.1 pu to 0.5 pu, causing a speed droop (Fig. 4.15(d)), that in turn forces a transition to TO mode (Mode 1). The rated i_{ds}^* is readily reestablished, and the speed controller increased $i_{qs}^{*'}$, bringing the speed back to its command value in less than 0.3 sec. Once the system reaches a steady-state condition, it switches back to EO mode (Mode 2) and a new search is initiated. Due to

Table 4.2 Induction machine parameters for efficiency optimization studies.

5 hp	230/460 V	13.4/6.7 A
4 poles	1710 rpm	NEMA class B

D^e-Q^e equivalent circuit parameters:

R_s=0.406 Ω	R_r=0.478 Ω
L_{ls}=0.00213 H	L_{lr}=0.00213 H
L_m=0.0494 H	

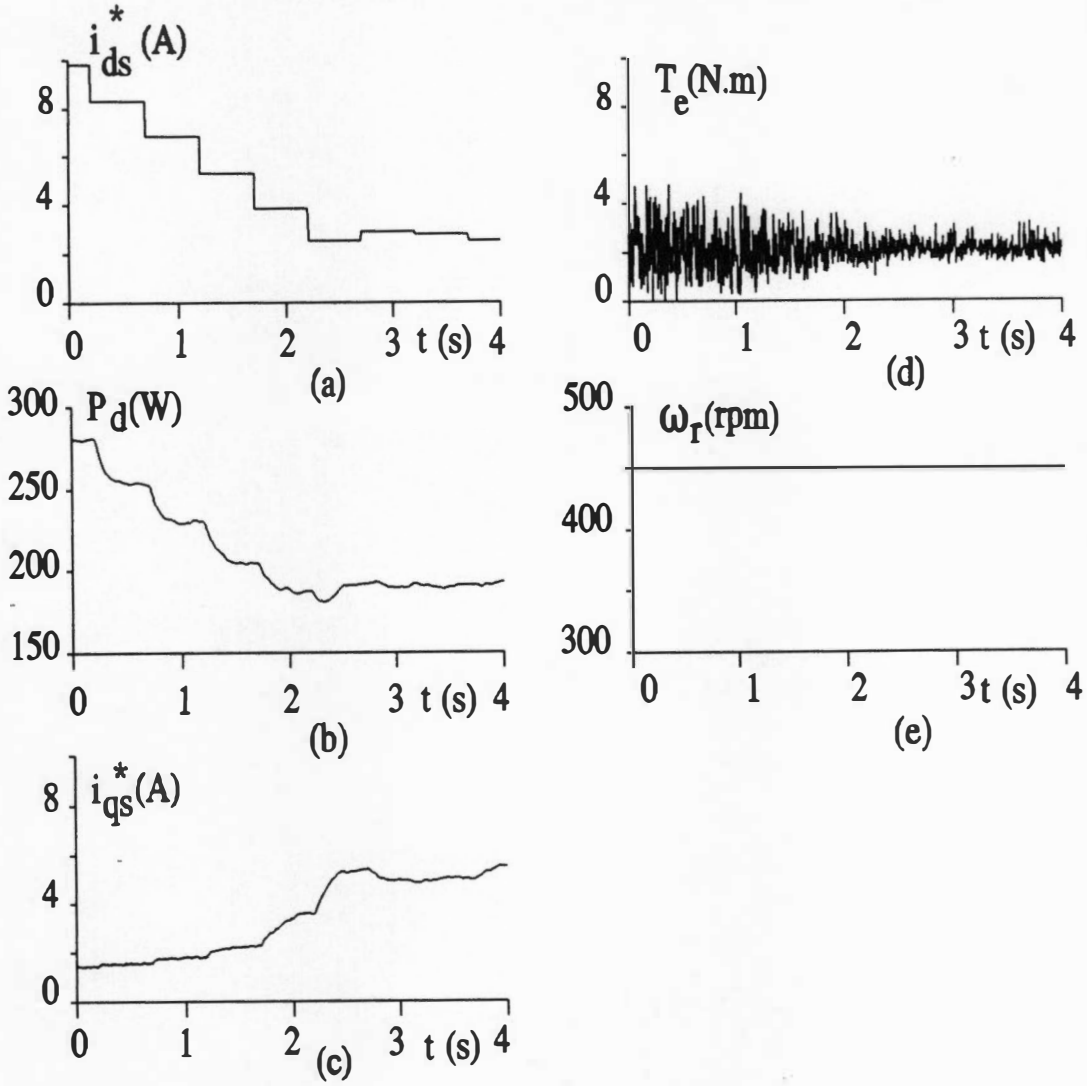


Fig. 4.13 Time domain simulated optimum efficiency search curves at ω_r=0.25 pu and T_L=0.1 pu. (a) Magnetizing current (i_{ds}^*). (b) DC link power (P_d). (c) Torque current (i_{qs}^*). (d) Machine torque (T_e). (e) Speed (ω_r)

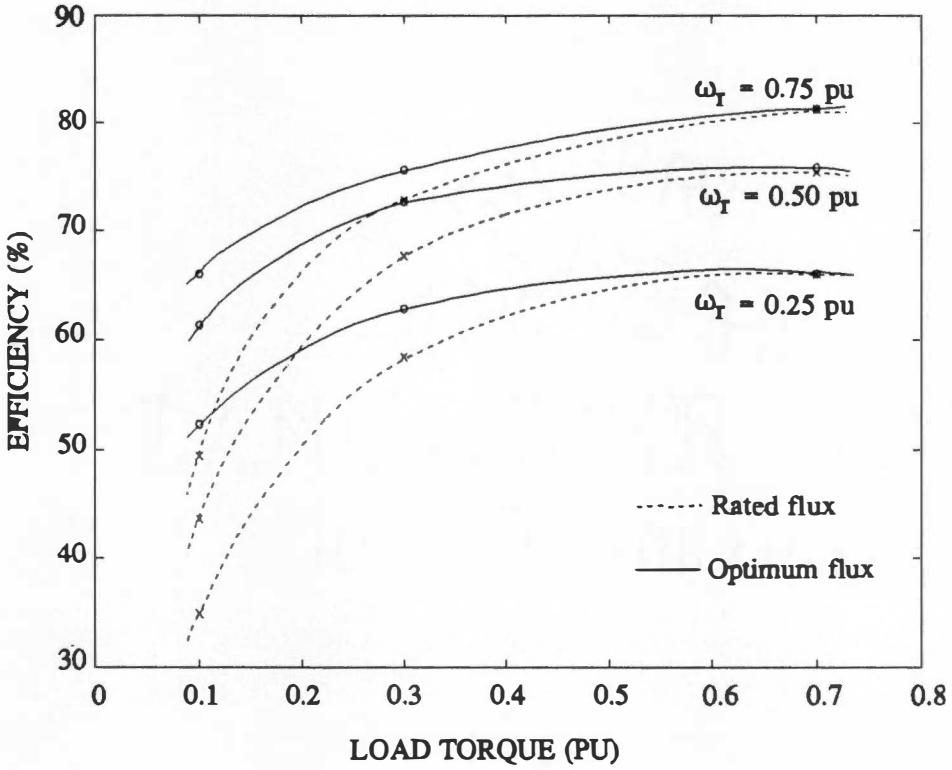


Fig. 4.14 Simulated efficiency curves.

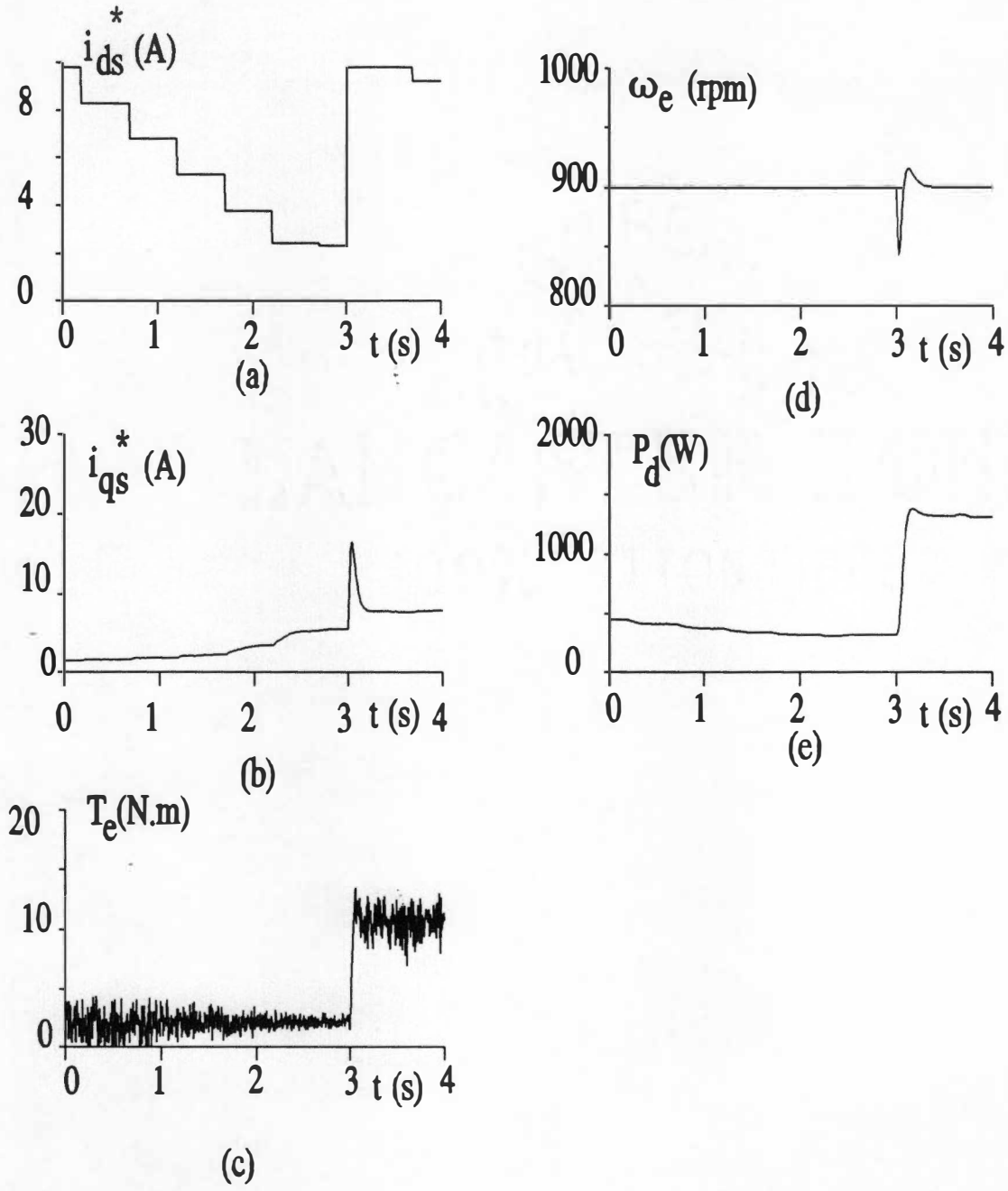


Fig. 4.15 Drive performance in time domain with sudden increase of load torque (transition to transient response optimization mode).

higher load torque the output gain I_b (in FEOPT) is reduced, what results in a smaller first step for the new search, as indicated in Fig. 4.15(a).

4.5 Hardware Circuit Design

The vector control system used in the experimental study was derived from an AC200 servo drive system, from General Electric Company. The modular servo system originally had the following components:

. Servo System Rack.

The rack is constructed with two receptacles to receive the plug in servo module and the power supply module. The rack has integrated cooling fan and DC link capacitor bank.

. Power Supply.

This module contains a power diode rectifier, rated at 230 V AC input, 325 V DC output. It also contains a low voltage, 80 KHz, auxiliary power supply for the control circuitry of the servo module. The control circuitry for dynamic breaking is also included, that automatically connects an external resistor across the DC bus, whenever the DC bus voltage reaches 385 Volts. In this way, the energy recovered from the deceleration of the drive is properly dumped, and dangerous DC voltage levels are prevented.

. Servo Module.

This module comes with a three-phase power transistor inverter and associated base drive and control circuit boards. The system originally used a sinusoidal

PWM current control, and had all the vector control functions implemented in dedicated hardware. As the system control board was missing, a new control board was designed.

The structure of the vector control system hardware is presented in Fig. 4.16. Its major component is a DSP-16 control board, from Ariel Corporation, based on Texas Instruments TMS320C25 DSP. The board is internally mounted on an IBM 286 compatible host computer, that serves as a development and debugging environment, as well as user interface during normal operation. The control board comes with two D/A channels, used to pass command currents i_a^*, i_b^* to the current controller, and two A/D channels, required for DC link voltage and current sensing. The speed interface is directly connected to the TMS320C25 data, address and control buses, and operates as a parallel peripheral of the DSP. The system is designed with basic protection and monitoring functions.

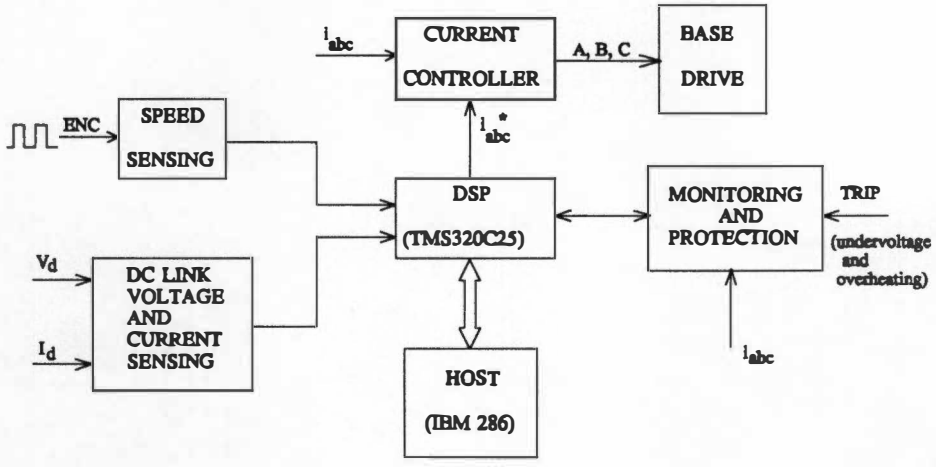


Fig. 4.16 Block diagram for vector control system hardware.

4.5.1 Current Controller

A hysteresis band current controller (HBPWM) was selected, because of its simplicity and robustness. Fig. 4.17 shows the circuit diagram of the proposed controller. Only two D/A channels are available; consequently, the reference current for phase c, i_c^*, is obtained by addition of i_a^* and i_b^*. After some very small filtering and scaling, the actual machine current i_a is subtracted from the command value i_a^*. The error is next compared to the hysteresis band to determine the state of phase leg A switches. Notice that, by using isolating diodes the hysteresis band for the three phases are controlled by a single trimpot. Identical circuits are used for phases b and c, and therefore, not shown here in detail.

4.5.2 Monitoring and Protection

Fig. 4.18 shows the circuit diagram for monitoring and protection. The system comes with built in inverter temperature sensor and under voltage protection, that are combined into a single trip signal (TRIP). Over current (OC) protection is obtained by extracting the positive and negative current profiles from the feedback current signals and comparing with pre-set limits. A small delay is introduced to prevent noise related false tripping. An emergence stop (E/S) push button is included for added safety. At power on, the delay circuits ensure a disable state for the drive circuit, and this must be cleared by the RESET push button. With all the protection circuitry enabled, a RUN command must be issued from the control software to effectively enable the base drive (STOP/RUN signal). LED lamps are associated with every trip/control signal for monitoring purpose.

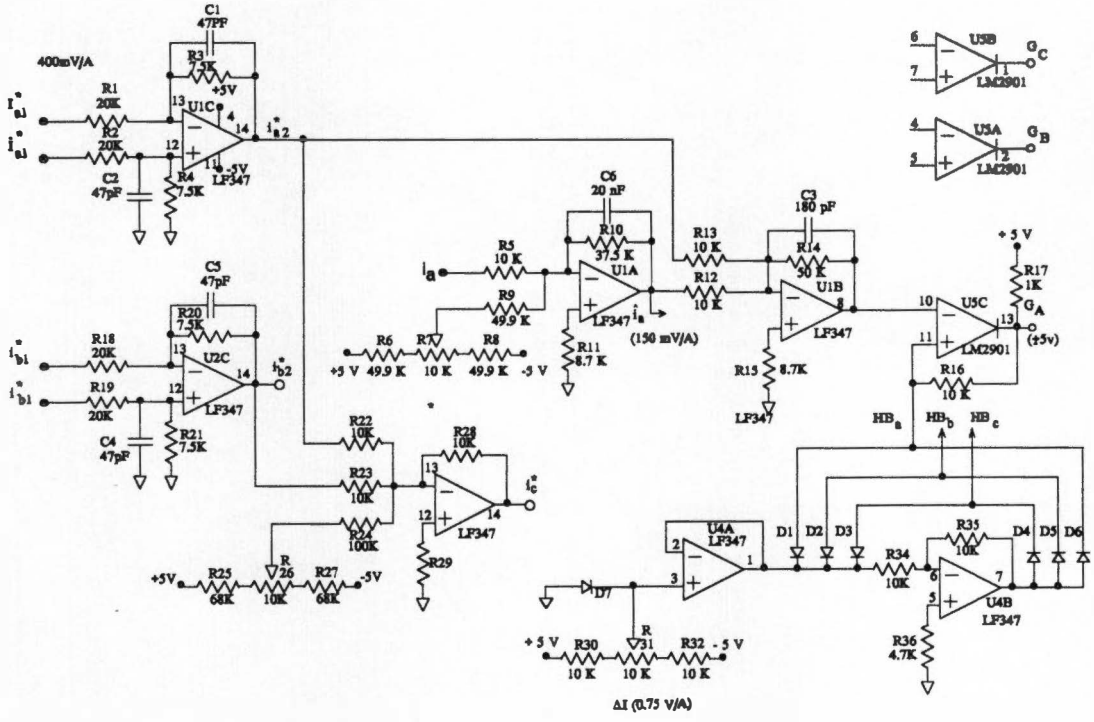


Fig. 4.17 Hysteresis band current controller.

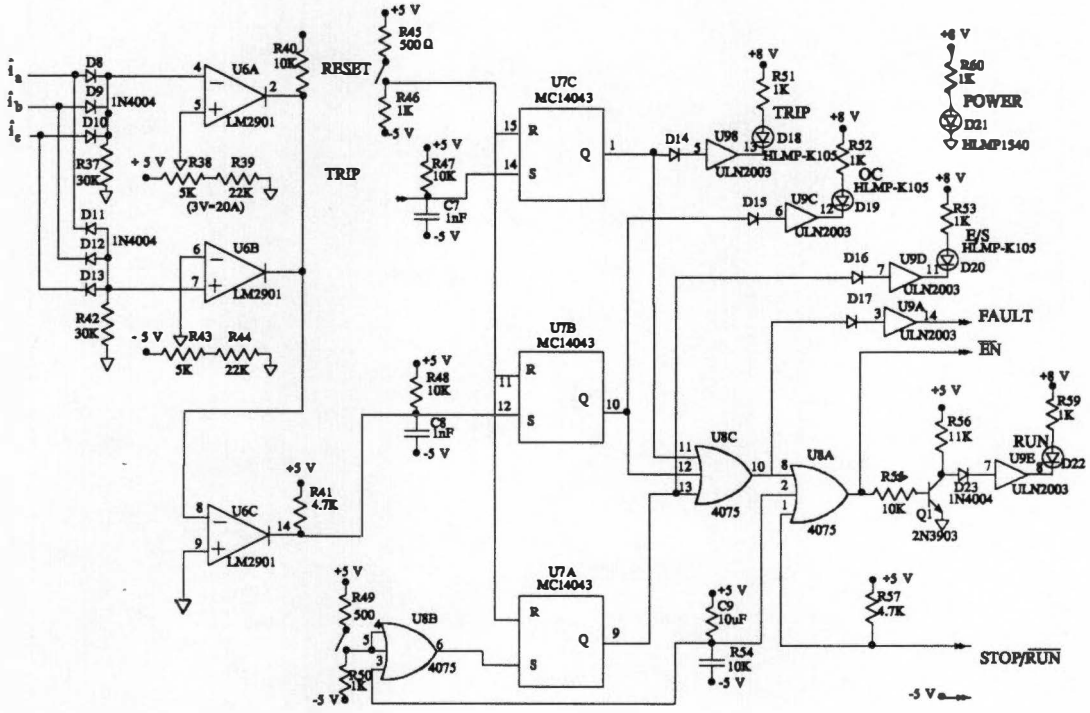


Fig. 4.18 Monitoring and protection circuit diagram.

4.5.3 Speed Measurement and Control Interface

Precise speed sensing is very important for proper operation of vector control systems. In fact, an error of 1% in speed measurement can result in severe degradation of drive performance.

The M/T method discussed in this section is characterized by very high precision in all speed range. Its implementation is shown in the circuit diagram of Fig. 4.19. Basically, it consists in counting pulses from an encoder (ENC) and from a high frequency clock (CLK). Speed information is obtained in software, by dividing the encoder counter increment (ΔM) by the clock counter increment (ΔT), with proper scaling, what will be explained later. Two 8-bit binary counters with output register (74ALS590) are used in cascade, for both ENC and CLK counters. The counter state is passed to the output register only at the positive edge of the register clock (RCK). By using the ENC pulse as the RCK signal of both counters, the contents of the output registers precisely reflects the actual speed.

A 3 to 8 decoder (74ALS138) is used to properly interface the counters to the DSP. Although the 74ALS590 output registers are directly connected to the DSP data bus, they are normally in a high impedance state, until the output enable pin (G) is asserted low by the 74ALS138 decoder. The output register content is then latched into the DSP.

The direction of rotation (DIR) and FAULT signals are similarly passed to the DSP through an octal bus transceiver (74ALS245). Finally, the START/STOP and counter clear CCLR signals are sent from the DSP to the control board through a dual 4-bit D flip-flop.

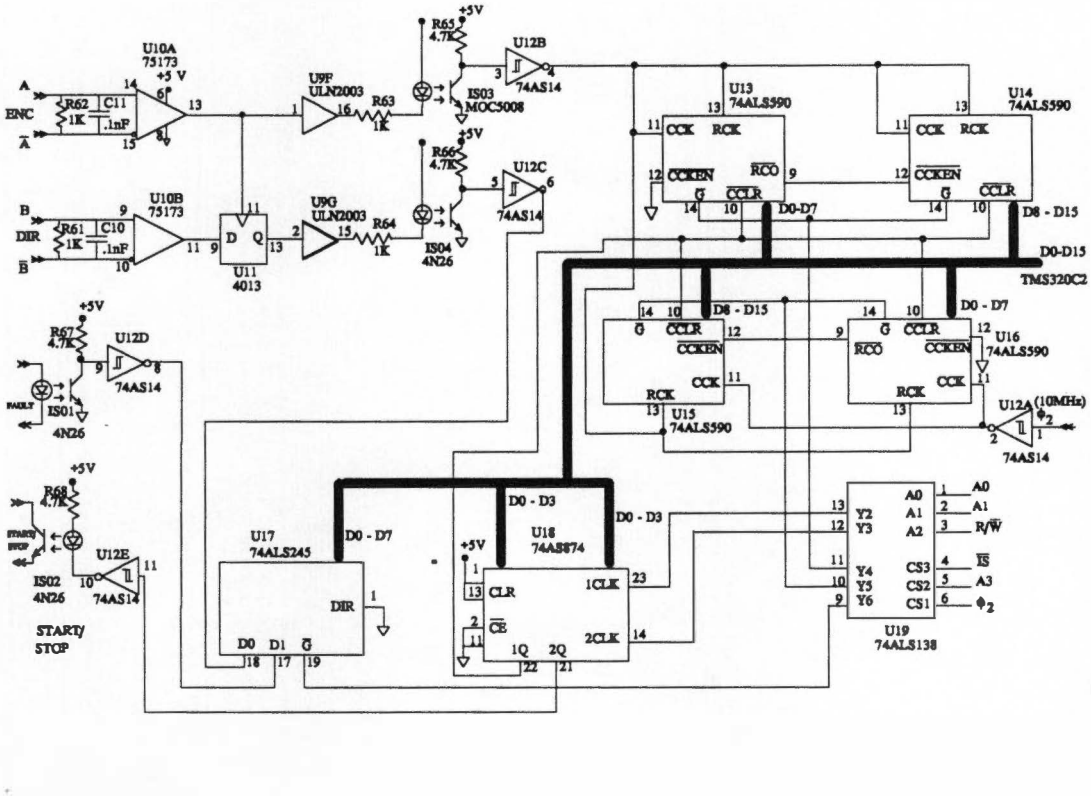


Fig. 4.19 Speed sensing and control interface.

4.5.4 DC Link Voltage and Current Sensing Interface

Adequate DC link voltage V_d and current I_d sensing is crucial for DC link power P_d computation, and consequently, for the correct operation of the efficiency controller.

DC link voltage sensing is shown in Fig. 4.20(a). A potential divider is initially used to bring the DC bus voltage to an adequate level and is next fed to an isolation amplifier (AM227), that ensures complete isolation between power and signal circuits. Its output is directly connected to the A/D input.

DC current sensing is performed by a high precision hall effect sensor (LEM LF 50-P), as indicated in Fig. 4.20(b). The sensor is placed between rectifier output and the DC capacitor bank. This location ensures that the PWM current harmonics generated by the inverter operation are filtered out by the DC link capacitor, resulting in an I_d waveform consisting of 6 pulses per period of 60 Hz supply. The LEM module works as a current transformer with a 1000:1 ratio, and with a 50Ω load resistor, a 50 mV/A gain is obtained. To ensure a better utilization of the A/D dynamic range, a low pass filter with a 3.81 gain is used. The reduction in DC link current bandwidth has no detrimental impact on P_d computation, since only its average value is relevant. In fact, as V_d is essentially constant, the average of P_d closely follows I_d average.

4.6 Real Time Software Design

The complete drive control system, including the fuzzy efficiency controller and pulsating torque compensator, was implemented in assembly language in a TMS320C25 DSP based control board from Ariel Corporation. Figure 4.21 shows the software structure

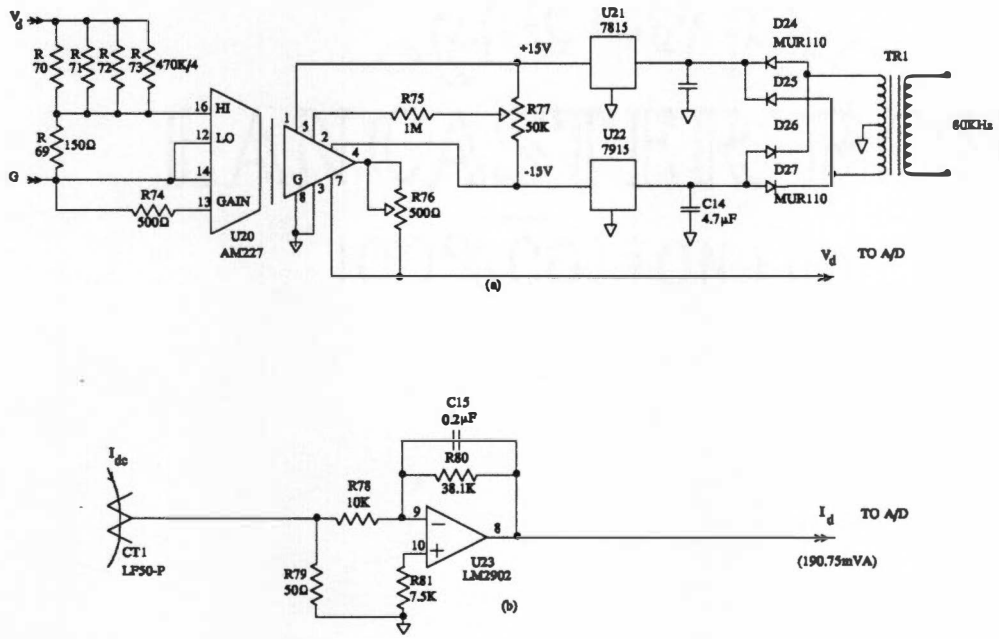


Fig. 4.20 DC link voltage and current sensing.

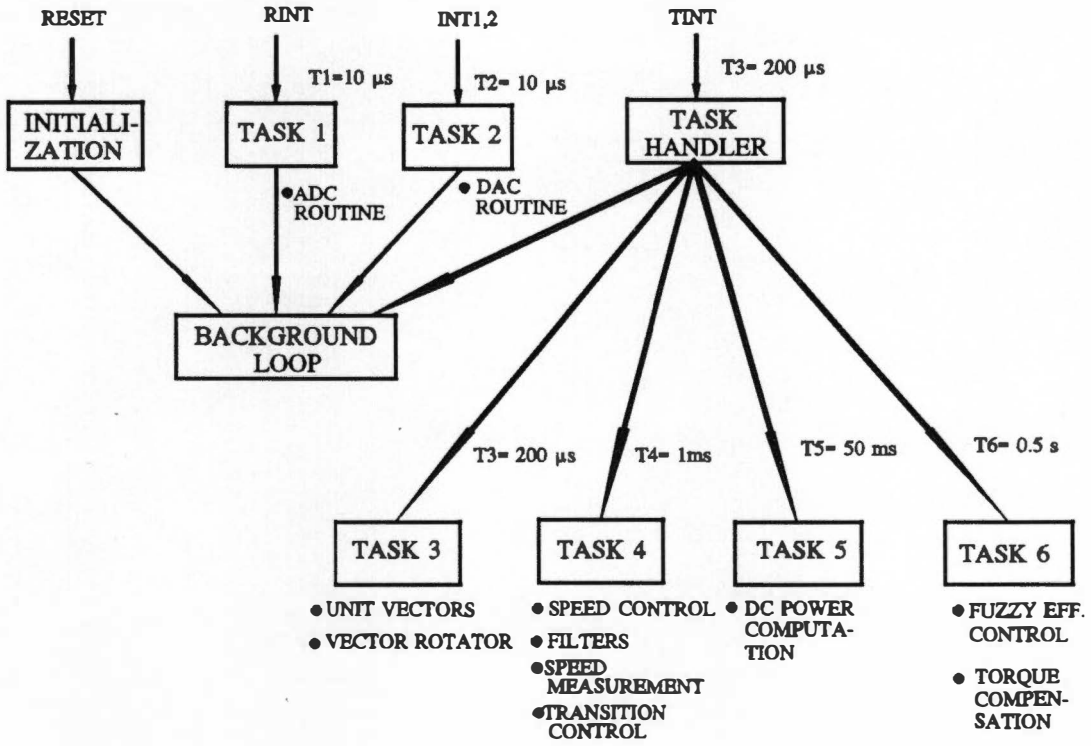


Fig. 4.21 Software structure of the drive system including the efficiency optimization controller

of the entire control system. Both the A/D and D/A converters operation are interrupt driven, and the maximum sampling frequency of 50 KHz per channel was selected, to benefit from the embedded 20 KHz bandwidth input and output filters.

The timer interrupt (TINT) was set to 200 μs, that is the sampling time of the vector control routines. The remaining routines are executed at lower sampling frequency, and controlled by software timers.

4.6.1 Basic Vector Control Functions

As previously indicated in the block diagram of Fig. 4.2, the implementation of the indirect vector control requires the generation of unit vectors sin θ_e and cos θ_e, from the speed signal ω_r and slip frequency ω_{sl}. The value of θ_e is obtained by discrete integration of ω_e, using double precision to ensure accurate computation at low speeds. A sine table with a resolution of 5.689 bits per degree was initially constructed and stored in program memory. The sin θ_e and cos θ_e values are obtained by simply scaling down the θ_e value to the interval [-1024, 1024] bits and then retrieving the proper value from the sine table (table look-up method). The unit vectors are next used to calculate the stationary reference currents i_{qs}^{s*} and i_{ds}^{s*}, from which the phase current references i_a^* and i_b^* are obtained. Phase C reference i_c^* is obtained from i_a^* and i_b^* by a summing operational amplifier (Fig.4.17).

4.6.2 Speed Computation

Speed computation will be discussed in details here. The basics of the M/T method have already being explained in the discussion of the speed interface hardware. The same principles are used here, as indicated by the speed computation flowchart of Fig. 4.22. The usual sampling time of 1 ms can be extended up to 5 ms, in case of very low speed. If no change occurs in the content of the encoder pulse counter (M) within the 5 ms extended period, the algorithm assumes $\omega_r = 0$. Under normal operation $M_{new} > M_{old}$ and the speed computation proceeds as indicated. For operation in both directions, the direction signal (DIR) must be probed, and the speed polarity corrected accordingly. Robust operation of the algorithm is achieved by imposing a pre-defined limit ($\Delta\omega_{max}$) to the speed change ($\Delta\omega_r = \omega_{new} - \omega_{old}$). This protects the system against eventual malfunction of the hardware interface or encoder failure. Furthermore, by selecting $\Delta\omega_{max}$ slightly larger than the maximum possible speed change under normal operation, no delay is introduced in the system, as would occur if a low pass filter were used.

4.6.3 Efficiency Optimization Controller

The real time fuzzy efficiency controller uses the same algorithm described in the simulation study. Consequently, only some implementation aspects will be discussed here.

The algorithm basically derives the next control action ($\Delta i_{ds}^*(k)$) from the measured change in input power ($\Delta P_d(k)$) and the last control action ($L\Delta i_{ds}^*(k)$). Initially, the interval index for each input variable is determined, thus defining the membership

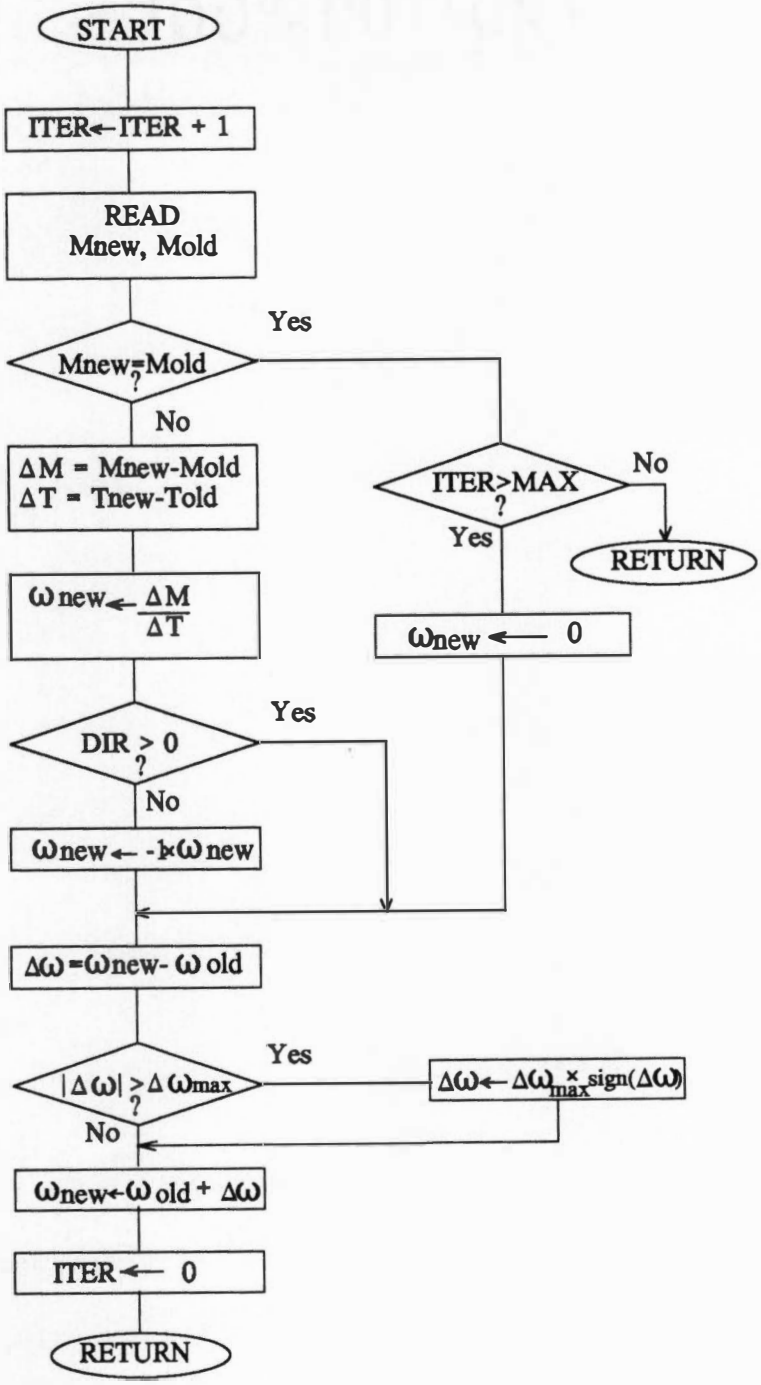


Fig. 4.22 Speed computation flowchart.

functions with non-zero degree of membership, as previously indicated in Fig. 4.11, for the case of ΔP_d.

Because the DSP does not have a division instruction, whenever possible a division operation is replaced by a multiplication. This approach is used in the computation of the degree of membership. For the particular value of $\Delta P_d(k)$ shown in Fig. 4.11, we have

$$\mu_{PM}(\Delta P_d) = (\Delta P_d(k) - P_3)\,(P_2 - P_3)^{-1} \tag{4.11}$$

As P_2 and P_3 are constants, $(P_2 - P_3)^{-1}$ is pre-computed and stored in data memory.

The rule base evaluation requires the retrieval of the control signal associated with every fired rule. This task is readily accomplished by the use of interval indices and a look up table containing the rule base. Let J be the index for ΔP_d , I the index for $L\Delta i_{ds}^*$, M and N the number of fuzzy sets for ΔP_d and $L\Delta i_{ds}^*$, respectively. The relative position (POS) of the first fired rule within the look up table is given by

$$POS = (J - 1)\,M + I \tag{4.12}$$

The second rule is located at POS+1, the third one at POS+N, and the last one at POS+N+1. Once the rule base evaluation is performed, the new control action is computed by the height defuzzification method.

4.7 Experimental Study

After careful software and hardware systems integration and debugging, the drive system was tested with a standard 5 hp NEMA class B induction machine, rated at 1710

rpm, 230 V, whose parameter were given on Table 4.2.

Typical AC supply phase voltage and current waveforms are shown in Fig. 4.23 for phase A, when the drive system was operating under rate flux, at a speed of 1350 rpm (0.75 pu) and a load torque of 7.68 lbf ft (0.5 pu). As discussed in Chapter 3, the diode rectifier conducts discontinuously, and therefore, significant amounts of harmonic currents are injected into the AC supply. The small unbalance in line voltages produces current pulses of varying amplitude, as shown. Some distortion in supply voltage waveform is also observable, generated by the harmonic current flow through the AC source impedance.

The machine input currents are shown in Fig. 4.24, when the drive was operating under rated flux, at 1350 rpm (0.75 pu) and delivering a load torque of 10.75 lbf ft (0.70 pu). The intrinsic operations of the hysteresis band PWM insures that the actual currents closely follow the sinusoidal command currents.

The operation of the fuzzy efficiency controller is illustrated in Fig. 4.25. The drive system was initially in a steady-state condition, at a speed ω_r = 450 rpm (0.25 pu) and load torque T_L = 1.54 lbf ft (0.10 pu). At t = 0 the efficiency controller was enabled, and the magnetizing current command i_{ds}^* was adaptively reduced on the basis of measured DC link power as shown in Fig. 4.25(a). The optimum operating point was achieved in only six steps. After that, the search algorithm remained active, with small Δi_{ds}^* steps, to ensure proper tracking in case the optimum point shifts due to parameter variation or slow changes in load torque. The operation of the pulsating torque compensation scheme is illustrated in Fig. 4.25(b). The bottom trace shows the

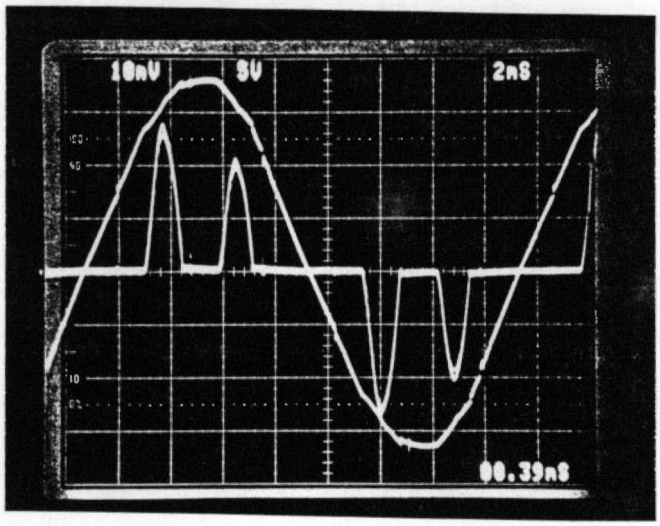


Fig. 4.23 AC supply phase voltage and current waveforms at ω_r=0.75 pu and T_L=0.50 pu. Scales: 50 V/div., 10 A/div., 2 msec./div.

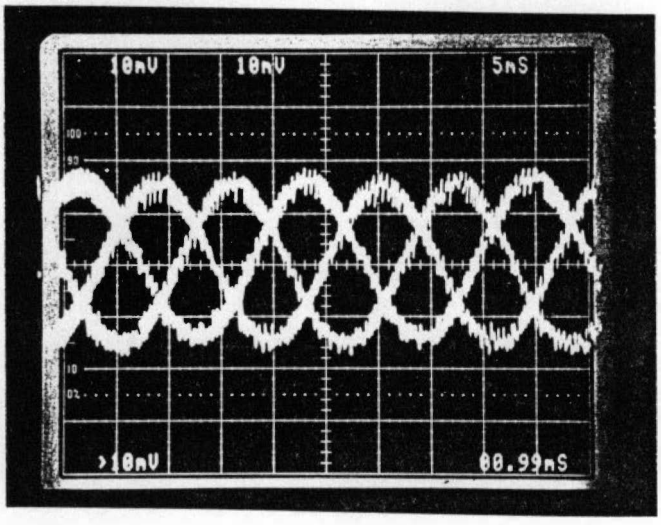


Fig. 4.24 Machine input currents at ω_r=0.75 pu and T_L=0.70 pu. Scales: 10 A/div., 2 msec./div.

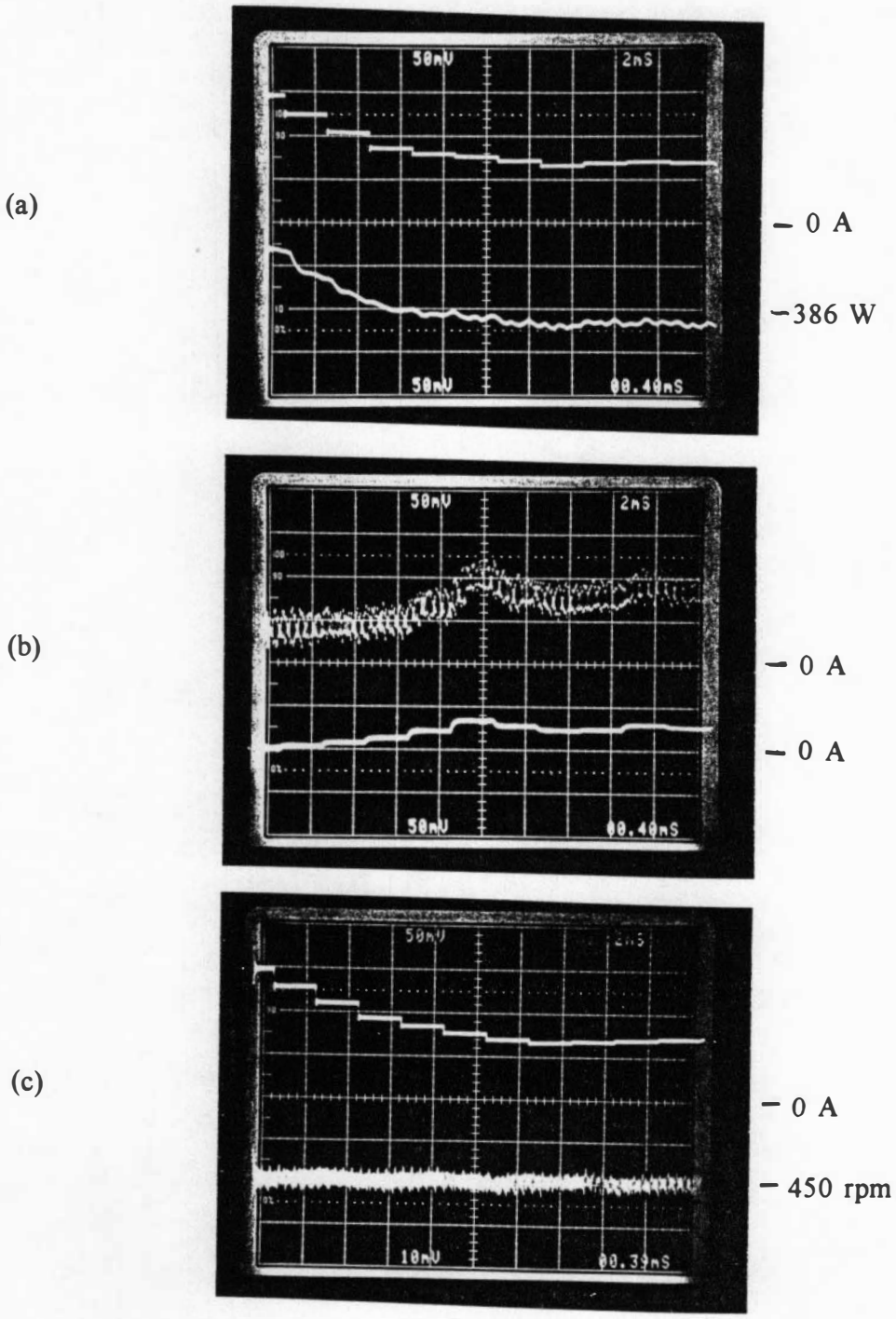


Fig. 4.25 Experimental search curve at ω_r=0.25 pu and T_L=0.10 pu.
(a) Top: i_{ds}^* (3.33 A/div.); Bottom: P_d (58.5 W/div.).
(b) Top: i_{qs}^* (1.67 A/div.); Bottom: $\Sigma\Delta i_{qs}^*$ (1.67 A/div.).
(c) Top: i_{ds}^* (3.33 A/div.); Bottom: ω_r- ω_r^* (61 rpm/div.).
Time scale: 2 sec./div.

compensating signal, $\Sigma\Delta i_{qs}^{*}$, whereas the upper trace represents the final torque current command, i_{qs}^{*}. It can be seeing that both signals follow similar profiles, demonstrating that the output of the speed controller ($i_{qs}^{*'}$) remains essentially constant. The effectiveness of the torque compensation can also be observed in Fig. 4.25(c), since the drive speed (bottom trace) remains practically constant.

Another search curves are shown in Fig. 4.26, for the case of ω_r = 900 rpm (0.5 pu) and T_L = 4.61 lbf ft (0.3 pu). As expected, the optimum i_{ds}^{*} is higher than that of Fig. 4.25(a), due to increased load torque. Accordingly, the observed reduction in DC link power (lower trace) is smaller, confirming that less efficiency gain is attainable as the load torque increases.

The associated torque current command, pulsating torque compensating signal and speed signal have similar profiles to those shown in Fig. 4.25 (b) and (c), and are not included here.

Some small fluctuations in DC link power are observable at any load condition, even at rated flux. These are caused by several factors, such as actual fluctuations in the load (dynamometer); the small unbalance in practical motors that creates negative sequence torques; and also some eventual small speed oscillations due to closed loop system dynamics. To reduce these fluctuations and ensure proper operation of the efficiency controller, a low pass filter of time constant τ = 0.5 sec. was introduced in the DC link power measurement routine, and the sampling interval for efficiency optimization increased to 2 sec.

To further demonstrate the effectiveness of the torque compensation scheme, the

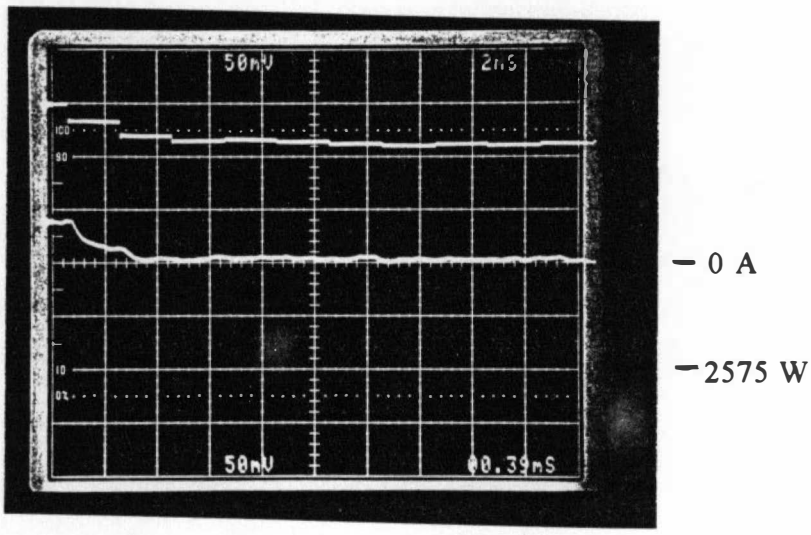


Fig. 4.26 Experimental search curve at ω_r=0.50 pu and T_L=0.30 pu.
Top: i_{ds}^* (3.33 A/div.); Bottom: P_d (58.5 W/div.)
Time scale: 2 sec./div.

same operating point of Fig. 4.25 is repeated in Fig. 4.27, but with the pulsating torque compensation disabled. The flux reduction and the attendant reduction in developed torque results in larger speed drops, as shown in Fig. 4.27(a). Of course, the speed controller reacts to restore the speed to its set value. However, as shown in Fig. 4.27(b) (bottom trace), the increase in torque current command i_{qs}^* is slow, and does not prevent the speed to drop. In some applications, speed fluctuations are not permissible, and in all cases they tend to create fluctuations in DC link power, that adversely affects the operation of the efficiency controller.

In practical drive system, the operating point can vary, either by changes in load torque or in command speed. Fig. 4.28 demonstrates the ability of the proposed control strategy to cope with sudden changes in operating condition. The drive was initially operating in a steady-state mode, at ω_r = 450 rpm (0.25 pu) and T_L = 1.54 lbf ft (0.10 pu), and the efficiency optimization controller brought i_{ds}^* (top trace) to its optimum value. At t=16 sec. the command speed ω_r^* was suddenly changed to 900 rpm, forcing the system to transition to a dynamic mode, where the rated i_{ds}^* was readily reestablished. At t = 19 sec. the system had already reached a new steady-state mode, as indicated by the speed signal (bottom trace in Fig. 4.28(a)). A new search for the optimum efficiency point was initiated, and when it was completed, another transient was imposed by setting ω_r^* = 450 rpm once more. In any case the drive speed response was quite fast, demonstrating the adequacy of the proposed method for applications where fast transient response must be maintained. Fig. 4.28(b) shows the torque current command i_{qs}^* (lower trace) for an identical ω_r^* profile, at same load condition. During the steady-state modes,

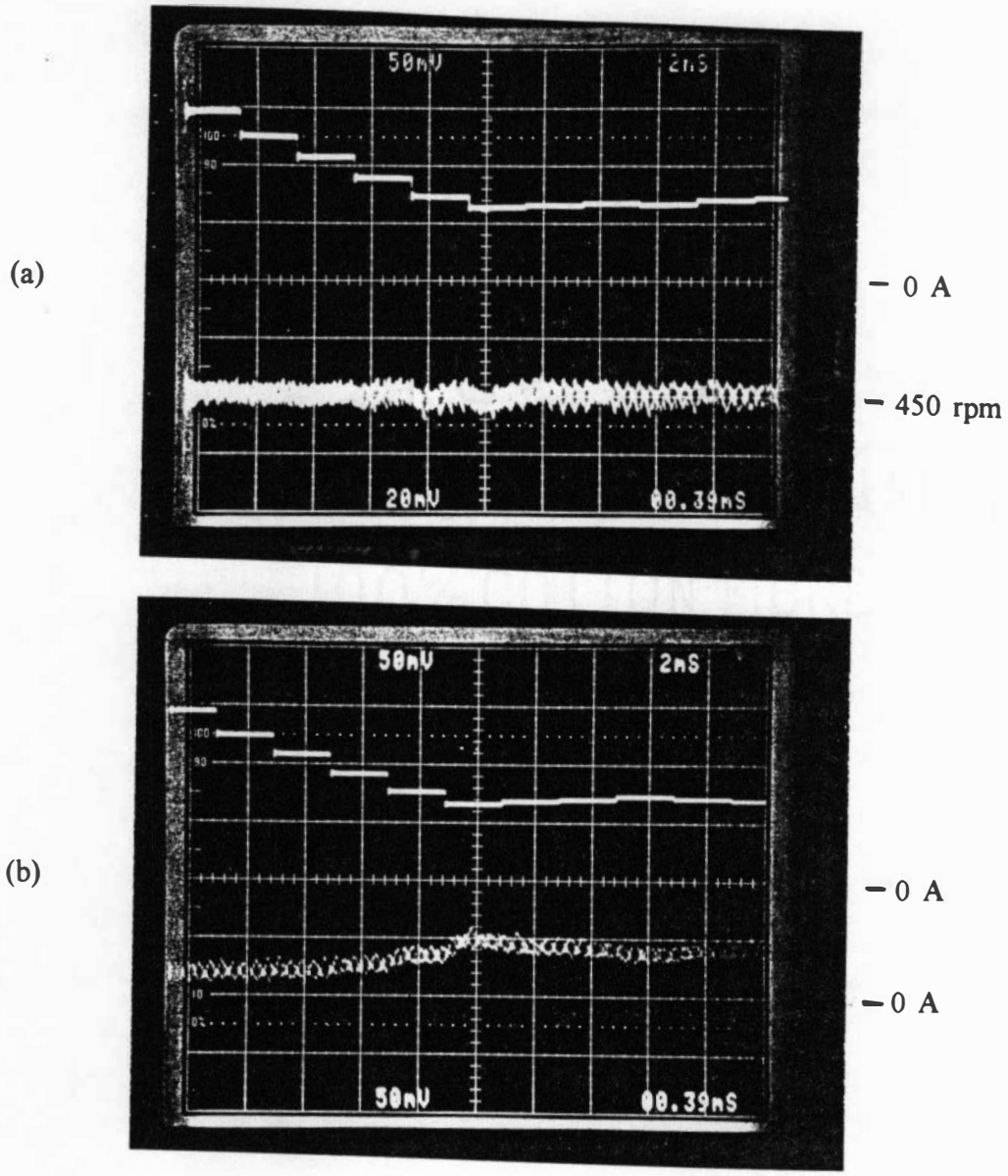


Fig. 4.27 Search curves at ω_r=0.25 pu and T_L=0.10 pu, without pulsating torque compensation.
(a) Top: i_{ds}^* (3.33 A/div.); Bottom: ω_r-ω_r^*, (61 rpm/div.)
(b) Top: i_{ds}^* (3.33 A/div.); Bottom: i_{qs}^* (3.33 A/div.).
Time scale: 2 sec./div.

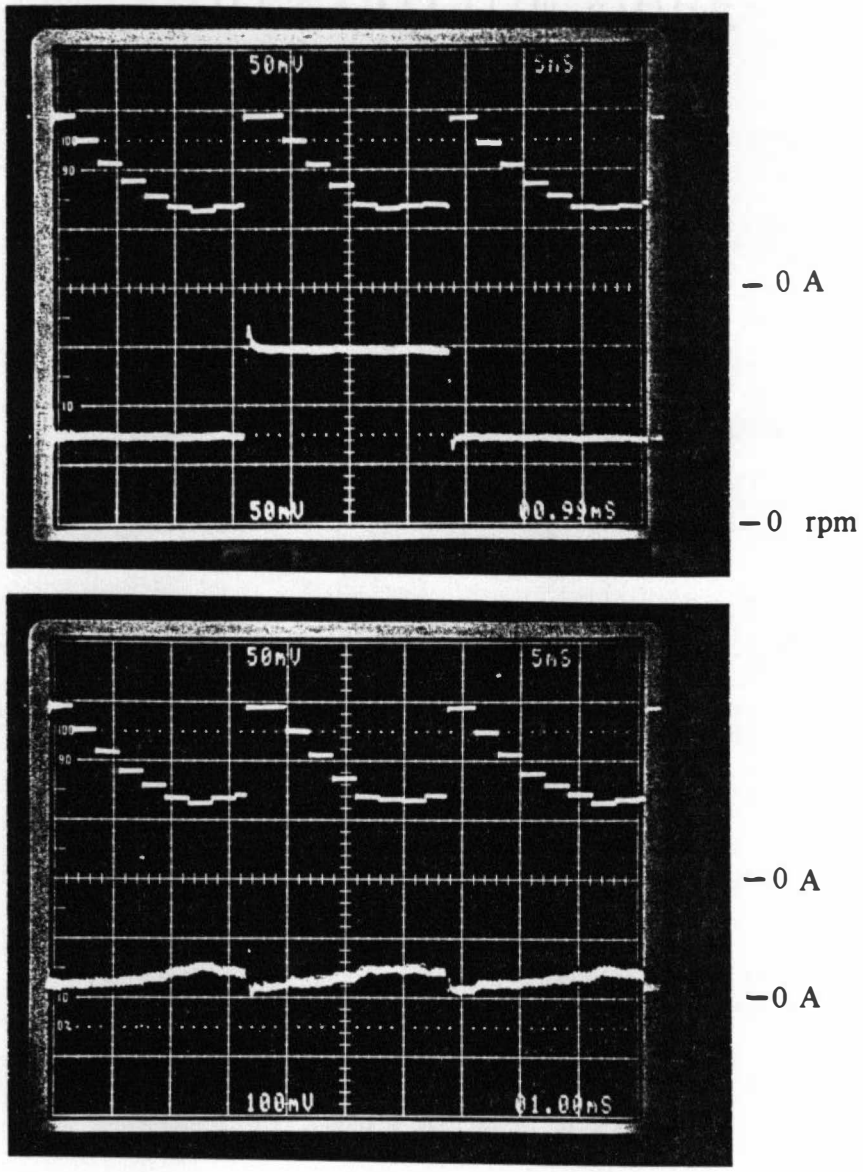


Fig. 4.28 Drive performance in time domain with sudden changes in command speed.
(a) Top: i_{ds}^* (3.33 A/div.); Bottom: ω_r (305 rpm/div.)
(b) Top: i_{ds}^* (3.33 A/div.); Bottom: i_{qs}^* (3.33 A/div.).
Time scale: 5 sec./div.

i_{qs}^* changes are mainly caused by the pulsating torque compensator, whereas in the transient modes they are solely governed by the speed controller, as indicated by the spikes in i_{qs}^* waveform.

The drive system was further tested at several operating points of the torque-speed plane, both at rated and optimum flux conditions, to assess the efficiency improvements. During experimentation, it was realized that the available power measurement equipment did not have adequate bandwidth to accurately measure the system input power. On the other hand, the DC link power measurement scheme operating at 50 KHz possesses very good precision. It was then clear that the system input power could be better estimated by computing the diode rectifier loss, and then adding to the measured DC link power. However, no specifications on the power diodes were available, and consequently, the diode volt-ampere (V-I) characteristic had to be obtained experimentally. The experimental data was fed to the software TABLE-CURVE [36] to obtain the diode voltage drop equation, similar to (3.49). At every operating point, the measured DC link power and voltage were used to compute the average DC link current ($I_{d(ave)}$), that in turn was substituted into (3.50) to obtain the rectifier loss.

The AC-200 system has a discharge resistor continuously connected across the DC link capacitor, such that no dangerous voltage levels remain after a few minutes the system has being shut down. However, this design feature creates extra power loss during normal operation. To make the results more general, this extra loss is deducted from the measured DC link power, prior to the diode rectifier loss and efficiency computations.

Finally, the overall system efficiency was determined, and resulting efficiency

curves are shown in Fig. 4.29. As mentioned earlier in this Chapter, the DC link power and the AC supply power follow similar profiles. Consequently, the efficiency curves of Fig. 4.29 reflect true optimum system operation. The efficiency gains are expressive, particularly at light load torques, but somewhat smaller than the predicted simulated efficiency gains. A number of factors can be enumerated to explain such differences. The most significant ones are:

. The experimental hysteresis-band size is larger than the one used in simulations, due to lock-out time effect. This translates in lower order harmonic voltages and consequently, extra harmonic losses in the machine;

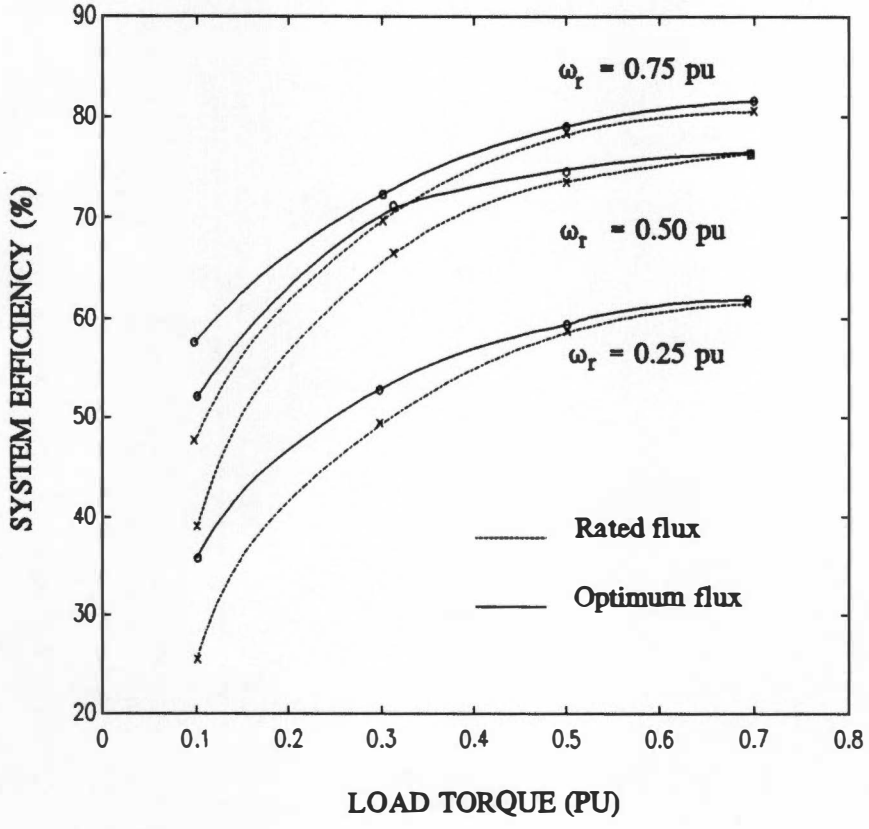


Fig. 4.29 Experimental efficiency curves.

. The 5 hp class B motor is not perfectly symmetrical, i.e., the flux distribution in the air-gap has a negative sequence component, that in turn creates a second harmonic torque. As a consequence, the resultant torque is smaller than that of a perfectly symmetrical machine, and this leads to lower efficiencies.

Irrespective of the practical limitations of both machine and converter systems, the proposed fuzzy controller is capable of achieving the maximum possible efficiency for a given operating condition.

The technique is currently being extended to an electric vehicle drive, as part of a technology transition project for Delco-Remy company. The drive system under investigation is somewhat similar to that shown in Fig. 4.2, except that the indirect vector control operates in a torque control mode, and the inverter is fed by a battery bank. Fuzzy efficiency control is engaged when the vehicle is traveling at constant speed, i.e., when the drive system is in a steady-state mode. The system switches back to a dynamic mode, if the torque command from the vehicle driver ($i_{qs}^{*'}$) is modified, or if it is detected that a speed variation has occurred. In such cases, rated flux is promptly restored to ensure maximum torque capability. The torque compensation scheme prevents torque pulsations during efficiency optimization, that could lead to reduction in riding comfort.

CHAPTER 5

FUZZY LOGIC BASED SLIP GAIN TUNING

5.1 Introduction

Indirect vector control is by far the most popular type of control for high performance induction motor drives. It offers the independent torque and flux control characteristics of a separately excited dc machine drive, with the advantages of lower machine cost, higher torque to inertia ratio and virtually maintenance free operation. One limitation of this control is its dependency on machine parameter, namely rotor resistance (R_r') and rotor inductance (L_r). It is the standard practice to use the nominal equivalent circuit parameters of the machine to compute the slip gain. For an unknown machine, the drive system can be self-commissioned, based on initial automated identification of parameters. However, machine parameters may vary widely in operating condition, due to changes in machine temperature and saturation, resulting in detuned operation.

Under detuned operation, there is a cross-coupling between flux and torque control loops, that adversely impacts their transient responses as well as results in incorrect steady state values of torque and flux. These effects will be illustrated with the help of the torque control system shown in Fig. 5.1. In this system, the flux command is kept constant at its rated value, while the torque command can be set to any value between

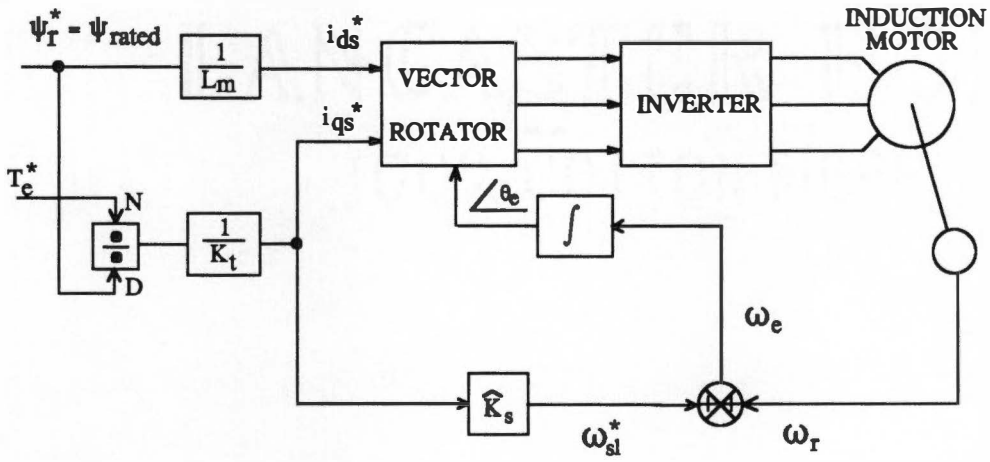


Fig.5.1 Indirect vector-controlled induction motor drive with open loop torque and flux control.

zero and the maximum torque. A fixed saturation factor is assumed, i.e., machine inductance are considered constant. The steady-state characteristics of developed torque and rotor flux are shown in Fig. 5.2, for three different values of slip gain estimate $\hat{K}_s$. When the slip gain estimate $\hat{K}_s$ precisely matches the actual value K_{so}, a linear torque transfer function is obtained, as shown by curve 2 of Fig. 5.2(a). Furthermore, the rotor flux ψ_r is unaffected by changes in torque, as indicated by curve 2 of Fig. 5.2(b). For the case $\hat{K}_s = 0.5\ K_{so}$ (curve 1), as the torque command is increased, the insufficient slip frequency causes the magnetizing current to increase. The associated overflux forces the machine into saturation and increases the core losses, that may result in overheating of the machine. On the other hand, for $\hat{K}_s = 2\ K_{so}$, the opposite situation occurs. The machine is demagnetized as the torque command increases, resulting in a reduction of the maximum torque capability, that in turn might cause the drive to stall at high load torques. It is also clear the loss of linearity of the input/output torque characteristic for $\hat{K}_s \neq K_{so}$.

For the case a closed loop torque control system, the torque error due to detuned operation would be corrected by action of the torque controller, such that at steady-state the torque demand would be met. However, this would require a higher stator current, that in turn would lead to extra copper losses and consequently, to a derating of the machine.

The transient effects of detuning are illustrated in Fig. 5.3. In Fig. 5.3(a) the slip gain was set to twice the correct value. The cross-coupling effect is quite evident, from higher order dynamics for torque and flux transients. The oscillatory torque response renders the system unsuitable for torque control applications. On the other hand, Fig.

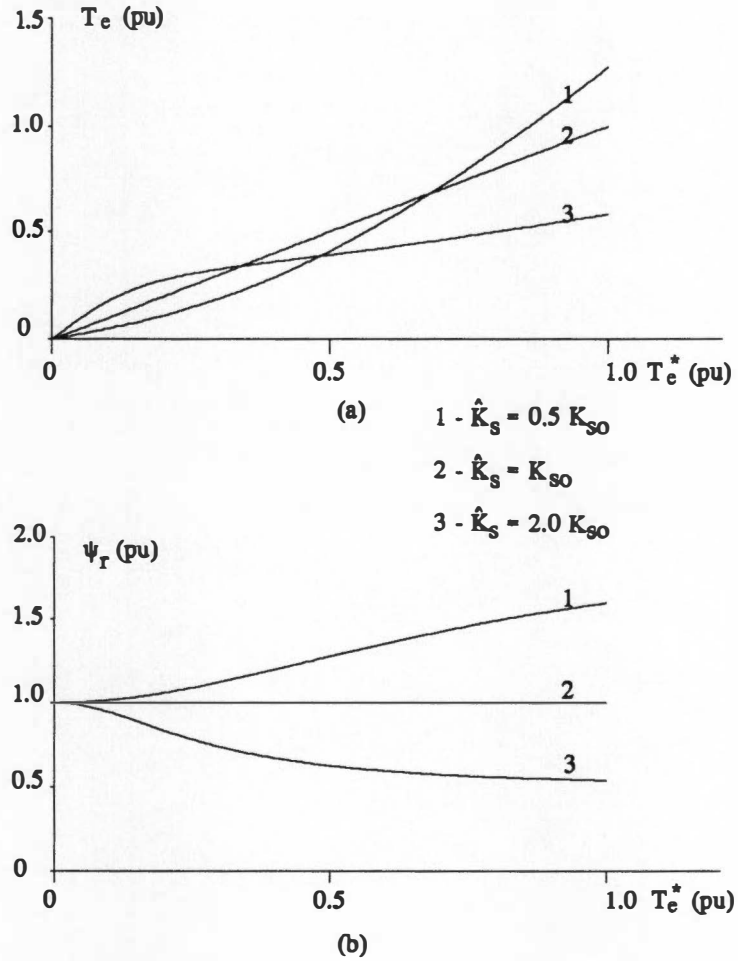


Fig. 5.2 Steady state detuning effects.
(a) Torque characteristics.
(b) Flux characteristics.

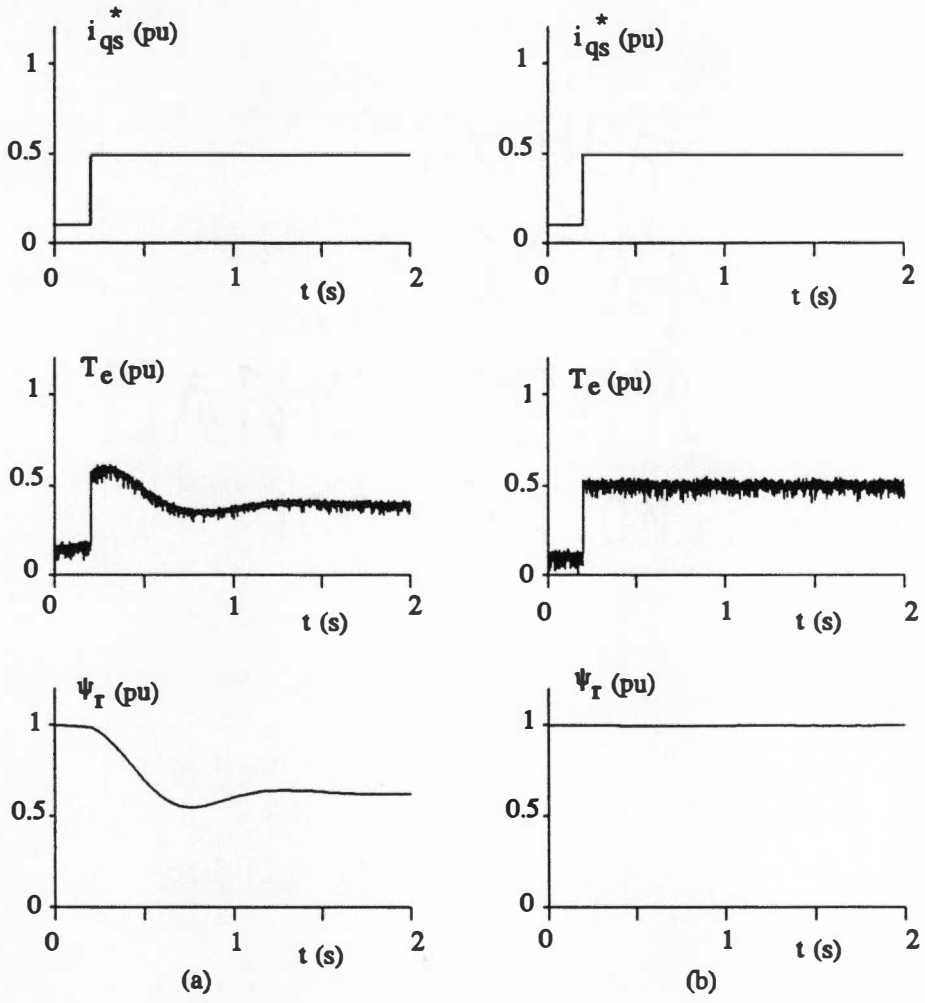


Fig. 5.3 Transient response for developed torque and rotor flux.
(a) Detuned condition ($K_s / K_{so} = 2.$)
(b) Tuned condition.

5.3(b) shows the ideal transient response for torque, as well as the perfectly decoupled flux response, for the case of correct slip gain.

Various on-line slip gain tuning techniques have been reported in the literature. Slip gain adaption based on the difference between ideal machine reactive power and actual reactive power was first reported by Garces [45]. Pseudo-random binary signal injection in one axis and then tuning on the basis of the induced signal in the other axis is described by Gabriel et al. [46]. Moreira et al [47] proposed a digital deadbeat controller, that derives the corrective term ΔK_s for the slip gain on the basis of an approximate machine inverse model and rotor flux estimates. The method, however, requires elaborated pre-computation of experimental saturation functions for air-gap flux and magnetizing inductance estimation. Zai and Lipo [50] used an extended Kalman filter to estimate the inverse rotor time constant on line, that makes use of the wide-band harmonic voltages, intrinsic to power inverters, as the random test signal. Systematic application of model-reference adaptive control (MRAC) technique based on reference torque model (ideal torque at tuned condition) and actual estimated torque for tuning the slip gain was reported in [48]. The MRAC tuning method was extended by Rowan et al. [49] to include voltage control method, which also evaluated the MRAC based torque (T_e), reactive power (Q), D-axis voltage (v_{ds}), Q -axis voltage (v_{qs}) and voltage magnitude (v_s) control methods, at fixed speed (60 Hz), and compared their performances. The study analyzed the advantages and limitations of all the methods, and concluded that none of them can solve the tuning problem satisfactorily in all the regions of torque-speed plane. Nevertheless, it was noticed from their study that the reactive power

model possesses good convergence characteristics, except at low torques. On the other hand, the D-axis voltage model exhibits excellent sensitivity to detuning, but its implementation at low speeds is complicated by the small signal to noise ratio, inherent to PWM inverters operating at small modulation factors. Furthermore, its dependency on stator resistance tends to reduce the method's precision at high load torques. Both model characteristics are discussed in detail in section 5.2. These complementary features suggests the creation of a new hybrid method, capable of performing slip gain tuning in practically the entire torque-speed plane. Such method must be able to select the most adequate model for each particular operating point, such that convergence to correct slip gain is achieved. The application of fuzzy logic in the derivation of this new method is discussed in the following sections. The results constitute the subject of a paper [63], to be presented at the 1993 international conference. on industrial electronics, control and instrumentation, IECON'93.

5.2 Fuzzy Tuning Controller

Fig. 5.4 shows the simplified block diagram of the proposed tuning controller, integrated with the indirect vector controlled drive system. The close loop speed controller generates the i_{qs}^* command which then multiplies with the slip gain $\hat{K}_s$ to generate the slip frequency command ω_{sl}^*. The MRAC system generates the reference models (A^*) with the help of command currents and frequency, compares with the estimated outputs of actual models (A), and then updates the slip gain $\hat{K}_s$ through the fuzzy tuning controller.

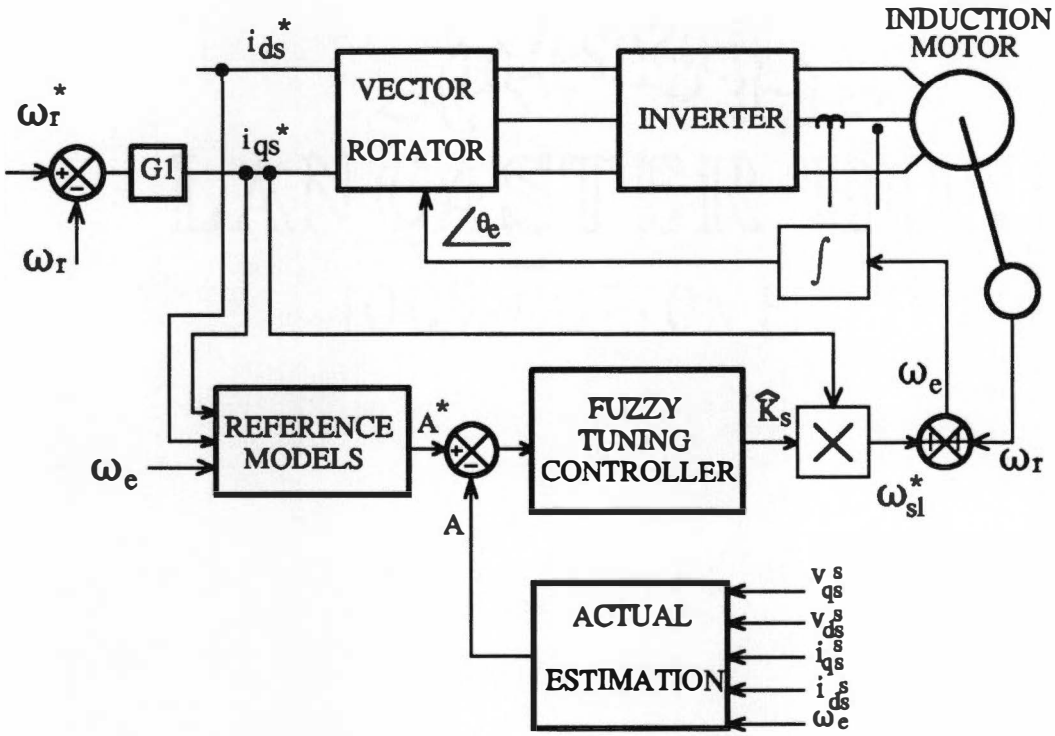


Fig. 5.4 Indirect vector-controlled induction motor drive showing the proposed fuzzy tuning controller.

The proposed method is actually based on two models, reactive power (Q) and D-axis voltage (v_{ds}), and consequently, A and A^* are vector variables.

5.2.1 Reactive Power and D-axis Voltage Regulators

The derivation of reference models for reactive power (Q^*) and D-axis voltage (v_{ds}^*) control was based on the synchronous frame D-Q model of the induction machine, shown in Fig. 3.3. The stator voltage equations are given by

$$v_{qs} = R_s i_{qs} + \frac{d}{dt}\psi_{qs} + \omega_e \psi_{ds} \tag{5.1}$$

$$v_{ds} = R_s i_{ds} + \frac{d}{dt}\psi_{ds} - \omega_e \psi_{qs} \tag{5.2}$$

Where the stator fluxes can be expressed as

$$\psi_{qs} = L_s i_{qs} + L_m i_{qr} \tag{5.3}$$

$$\psi_{ds} = L_s i_{ds} + L_m i_{dr} \tag{5.4}$$

Under steady-state field orientation the following conditions apply [17]

$$\psi_{qr} = L_r i_{qr} + L_m i_{qs} = 0 \tag{5.5}$$

$$i_{dr} = 0 \tag{5.6}$$

From equation (5.5), i_{qr} can be expressed as

$$i_{qr} = -\frac{L_m}{L_r} i_{qs} \tag{5.7}$$

Substitution of (5.6) and (5.7) into stator flux equations 5.3 and 5.4 leads to

$$\psi_{qs} = (L_s - \frac{L_m^2}{L_r}) i_{qs} = L_o i_{qs} \tag{5.8}$$

$$\psi_{ds} = L_s i_{ds} \tag{5.9}$$

Finally, substitution of (5.8) and (5.9), and the steady-state conditions of equation 5.10, into (5.1) and (5.2), results in the field oriented stator voltage equations 5.11 and 5.12.

$$\frac{d}{dt}\psi_{qs} = \frac{d}{dt}\psi_{ds} = 0 \tag{5.10}$$

$$v_{qs} = R_s i_{qs} + \omega_e L_s i_{ds} \tag{5.11}$$

$$v_{ds} = R_s i_{ds} - \omega_e L_o i_{qs} \tag{5.12}$$

The steady-state reactive power of the machine [49] is expressed by

$$Q = v_{qs} i_{ds} - v_{ds} i_{qs} \tag{5.13}$$

Substituting v_{qs} and v_{ds} from (5.11) and (5.12) into equation 5.13 leads to

$$Q = \omega_e (L_s i_{ds}^2 + L_o i_{qs}^2) \tag{5.14}$$

The reference quantities Q^* and v_{ds}^* are readily obtained from the field oriented equations (5.12) and (5.14) respectively as

$$v_{ds}^{*} = \hat{R}_s i_{ds}^{*} - \omega_e \hat{L}_\sigma i_{qs}^{*} \tag{5.15}$$

$$Q^{*} = \omega_e (\hat{L}_s i_{ds}^{*2} + \hat{L}_\sigma i_{qs}^{*2}) \tag{5.16}$$

Estimates of L_s, L_σ and R_s are required, but no rotor parameter is involved. The feedback variables Q is computed from (5.13), and actual v_{ds} is given by

$$v_{ds} = v_{qs}^{s} \sin\theta_e + v_{ds}^{s} \cos\theta_e \tag{5.17}$$

where $\cos\theta_e$ and $\sin\theta_e$ are the unit vectors available in the vector controller.

5.2.2 Derivation of Combined Error Signal

Fig. 5.5 shows the details of the proposed control system. The scheme depends on reference model computation of reactive power (Q^{*}) and D - axis voltage (v_{ds}^{*}) as discussed above. The reference models are compared with the respective actual estimates of the quantities, and the corresponding loop error is divided by a base value to convert to per unit variables. Handling the variables in pu form is convenient for fuzzy controller. Besides, the same controller can be easily extended to other drive systems. The base values for the respective variables are essentially the same as reference variables, i.e., $Q_b = Q^{*} + \varepsilon_1$ and $V_b = |\omega_e L_\sigma i_{qs}^{*}| + \varepsilon_2$, where the parameters epsilon (a small positive constants) have been added to avoid indeterminate computation, in case the reference model approaches zero at critical conditions. Note that the reverse polarity of Q is due to opposite behavior of v_{ds} and Q with respect to $\hat{K}s$, which is explained later. The first fuzzy logic controller (FLC-1) generates the weighting factor K_f, which permits

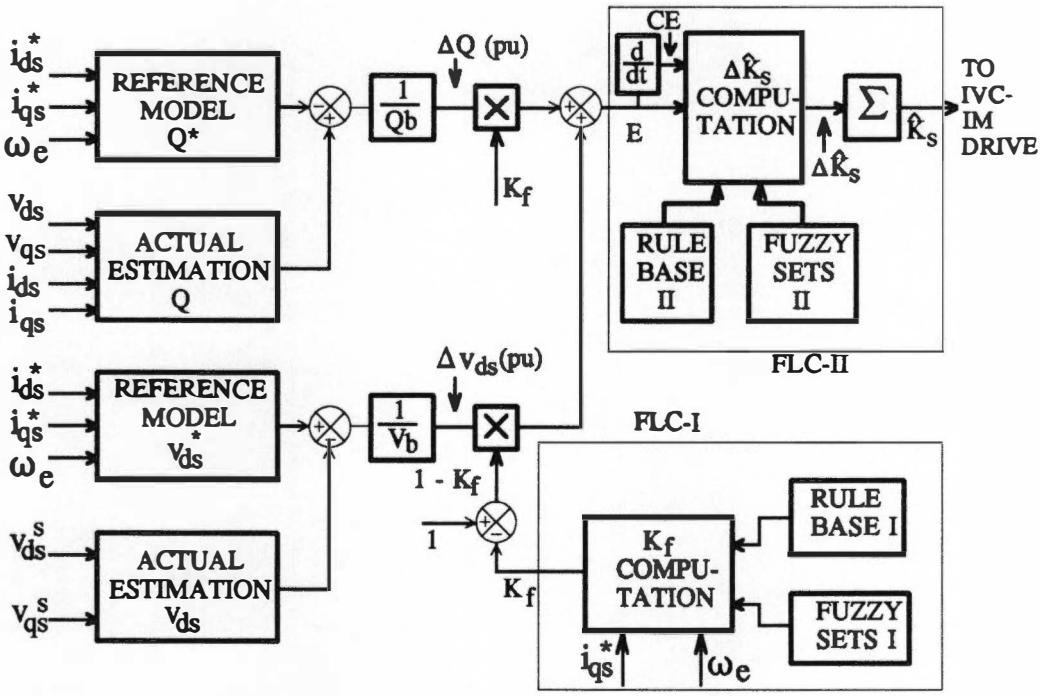


Fig. 5.5 Fuzzy logic based MRAC tuning control block diagram.

appropriate distribution of Q control and v_{ds} control on the iqs^*-ω_e, i.e., torque-speed plane. The combined error signal is defined as

$$E = \Delta Q\, K_f + \Delta v_{ds}\,(1 - K_f) \tag{5.18}$$

5.2.3 Design of the Fuzzy Tuning Controller

The second fuzzy controller (FLC-2) generates the corrective incremental slip gain $\Delta\hat{K}_s$, as shown in Fig. 5.5. As the error signal is derived from two different models, both its steady-state and dynamic characteristics vary with the operating point. Under such conditions, the use of a fuzzy controller, that inherently implements distinct control laws at different operating points, ensures fast convergence at practically any point in the torque-speed plane.

In ideally tuned condition, the signals ΔQ and Δv_{ds}, and correspondingly the E signal, will be zero and the slip gain $\hat{K}_s$ will be set to the correct value K_{so}. If the system is detuned, for example due to change of rotor resistance, the actual Q and v_{ds} variables will deviate from the respective reference variables. The resulting error will alter the $\hat{K}_s$ value until the system becomes tuned, i.e., E = 0. In order to derive the knowledge required for the design of the fuzzy controllers, the static characteristics of both Q and v_{ds} loops were initially investigated through simulation. To this end, the speed control system of Fig. 5.4 was used, but with tuning algorithm disabled. The slip gain $\hat{K}_s$ was intentionally varied from 0.5 to 2 times the correct value K_{so}. The speed control loop ensured any steady-state operating point on torque-speed plane, even with detuned condition. Fig. 5.6 shows the normalized error relation with normalized slip gain ($\hat{K}_s/K_{so}$)

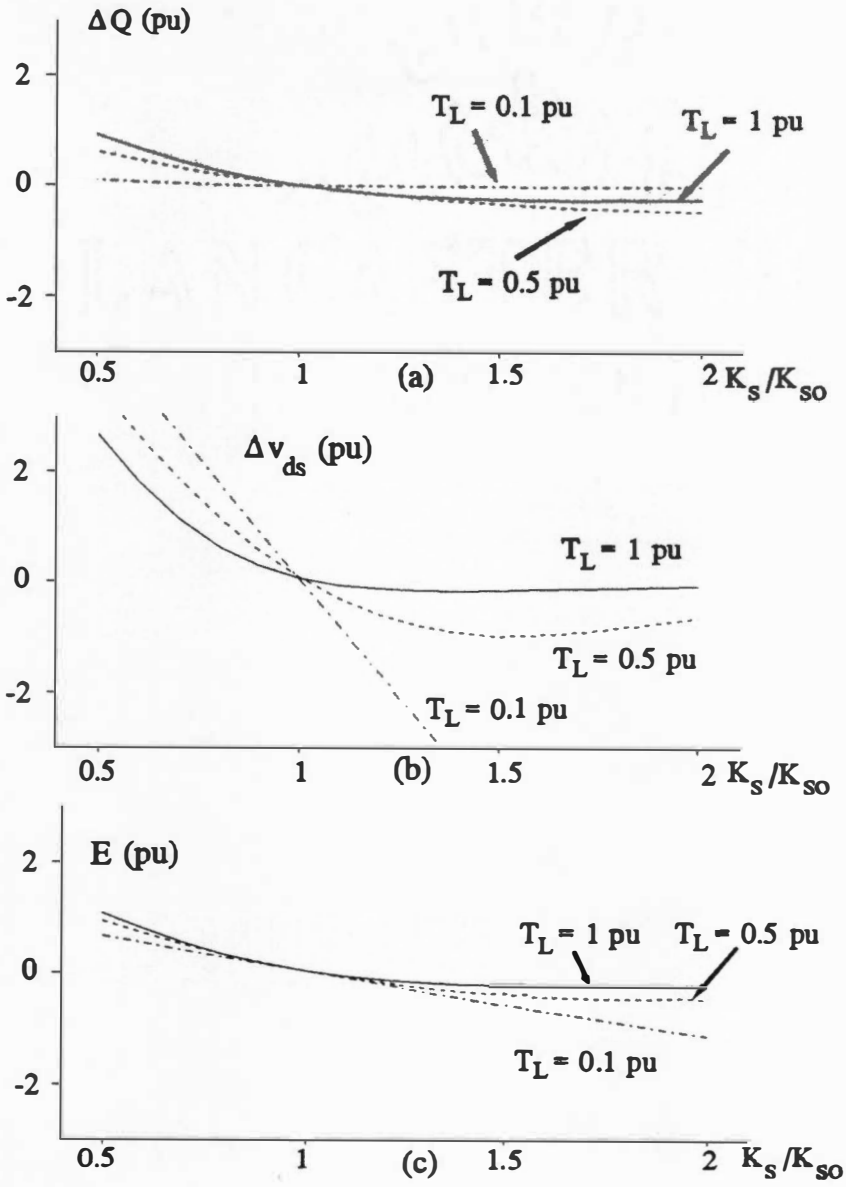


Fig. 5.6 Normalized control loop error vs. normalized slip gain curves (ω_r=0.5 pu).

(a) Reactive power error.

(b) D-axis voltage error.

(c) Combined error.

at speed ω_r = 0.5 pu, for different load torque conditions. Note that, at ideally tuned condition $\hat{K}_s/K_{so}$ = 1, and correspondingly ΔQ (pu) = Δv_{ds}(pu) = E(pu) = 0. If the system is detuned, for example with lower $\hat{K}_s$ (i.e., higher rotor resistance), then $Q > Q^*$, and the controller will raise $\hat{K}_s$ until the tuning point is reached. Note that the loop error tends to be very small at low torque, but increases with a higher load torque. Fig. 5.6(b) shows similar curves for the voltage control loop. The voltage loop error tends to be very large at detuned condition, except at T_L = 1.0 pu and $\hat{K}_s/K_{so} > 1$. These characteristics indicate that the parameter K_f should be large at high torque region, but small at low torque region.

Consider now the parameter variation problem in (5.15) and (5.16). If the machine parameters deviate from those used in the reference model, the slip gain will be tuned to an incorrect value. The inductance L_s will decrease by saturation caused by air-gap flux, but its value will remain unchanged if rated flux is maintained. On the other hand, the inductance L_σ will vary by saturation due to stator current, i.e., is load dependant. The stator resistance in (5.16) will vary with stator temperature, but its effect can be negligible at high speed. The parameter variation effect, along with the characteristics in Fig. 5.6(a) and (b), indicate that the v_{ds} control method is better at high speed, low torque region, whereas Q control method is superior at low speed high torque region.

Fig. 5.6(c) shows the combined error E (pu) as a function of $\hat{K}_s/K_{so}$, with optimum K_f distribution on torque-speed plane, obtained through FLC1. Fig. 5.7 gives the membership functions for frequency (i.e., speed), the current i_{qs}^* and the weighting factor

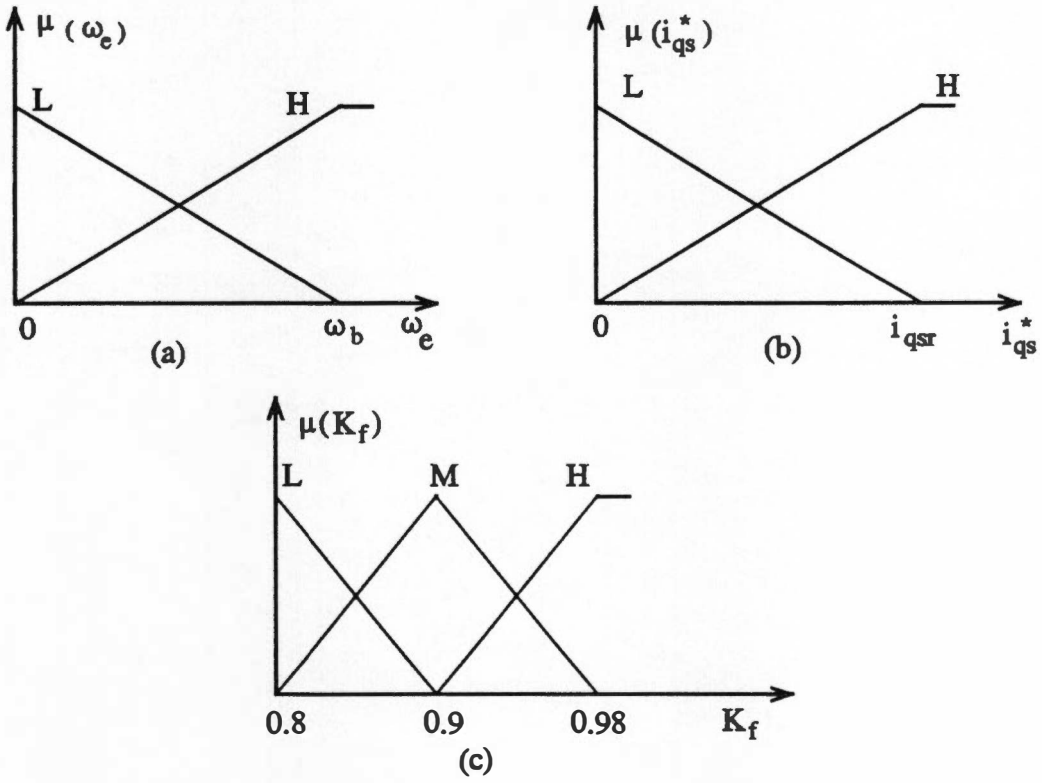


Fig. 5.7 Membership functions for FLC1.
(a) Speed (ω_e)
(b) Torque current (i_{qs}^*).
(c) Weighting factor (K_f).

K_f, and Table 5.1 gives the corresponding rule base for fuzzy controller FLC-1. An example rule can be stated as

IF speed (ω_e) is low (L) and torque (i_{qs}^*) is high (H) THEN weighting factor (K_f) is high (H).

In fuzzy control, an imprecise information can be useful. For the controller FLC-1, even if the system is detuned, i_{qs}^* can still be used as a measure of torque, at approximately rated flux condition. As mentioned before, the control strategy built into Rule Base I ensures high sensitivity to detuning for any given operating condition.

Fig. 5.8 gives the membership functions for error (E), change in error (CE) and slip gain increment ($\Delta\hat{K}_s$) for the controller FLC-II, where input and output gains were used to normalize the variables in the [-1,1] interval.

Table 5.1 Rule base for weighting factor (K_f) calculation.

i_{qs}^* \ ω_e	H	L
H	M	H
L	L	M

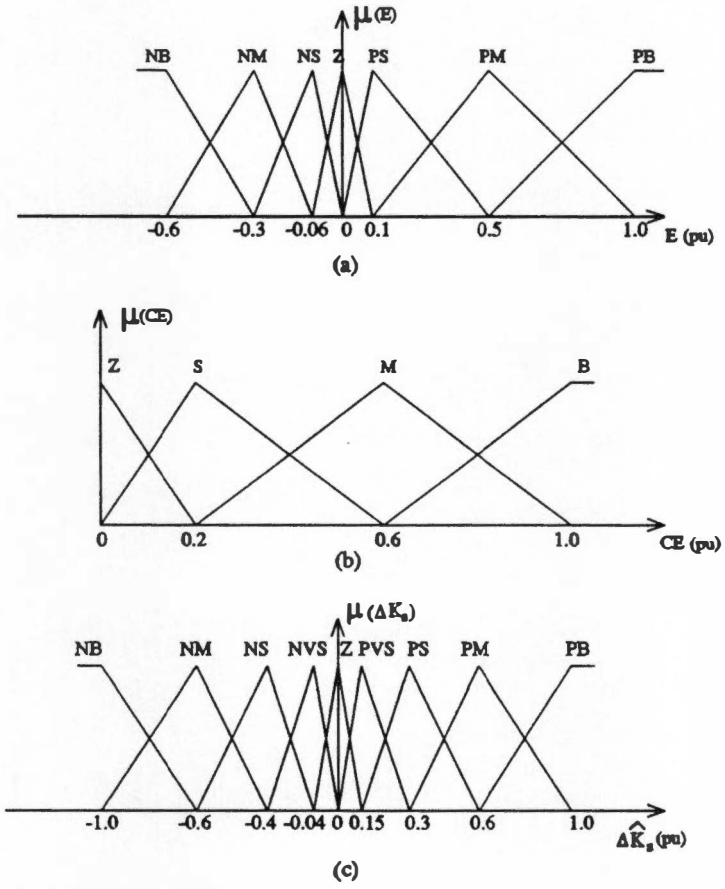


Fig. 5.8 Membership functions for FLC-2.
(a) Error (E).
(b) Change in error (CE).
(c) Increment in slip gain ($\Delta\hat{K}_s$)

The asymmetry of the functions gives better resolution in control, as the respective variable approaches zero. The control loop was designed to provide overdamped $\dot{K}_s$ response, and therefore, only CE magnitude was considered, resulting in simplification of the rule base, as shown on Table 5.2. Broadly speaking, the control strategy, embedded into rule base 2, consists in selecting a ΔK_s increment that forces the combined error E to approach zero, at a desired rate CE. The final rule base is a result of a fine tuning process, that used intensive simulation of the drive system, under various operating points in the torque-speed plane.

5.3 Simulation Study

Once the control algorithm was formulated in detail, the complete drive system with the proposed controller, as shown in Figs. 5.4 and 5.5, was simulated in SIMNON. The system has an outer speed control loop and a hysteresis-band PWM (HBPWM) in the inner loop. Simulation studies were conducted using 5 hp induction servomotor, originally used in the General Electric AC-200 vector control system, whose parameters are given on Table 5.3.

In order to keep the simulation time small, ideal (lossless) models were used for the rectifier and inverter subsystems, and a standard synchronous frame D-Q model was selected for the machine. The fuzzy controller was implemented using the same methodology described for the fuzzy efficiency controller, but here a 1 ms sampling time was selected. All SIMNON routines are given in Appendix A.3.

Table 5.2 Rule base for increment of slip gain (ΔK_s)

CE \ E	NB	NM	NS	ZE	PS	PM	PB
B	NM	NS	ZE	ZE	PVS	PM	PS
M	NB	NM	ZE	ZE	PVS	PM	PS
S	NB	NM	NVS	ZE	PVS	PB	PM
Z	NB	NB	NM	ZE	PVS	PB	PB

Table 5.3 Parameters for the induction servomotor.

5 hp 180 V 15 A		
4 poles 2200 rpm		
D^e-Q^e equivalent circuit parameters		
R_s=0.177 Ω	R_r=0.318 Ω	R_m=121.4 Ω
L_{ls}=0.00205 H	L_{lr}=0.00130 H	L_m=0.0328 H

As the system dynamics varies with the slip gain, performance studies were initially carried out to fine tune the membership functions as well as the rule bases of both controllers, until optimum performance was obtained in the tuning process. Fig.5.9 shows the normalized slip gain ($\hat{K}_s/K_{so}$) and error E(pu) versus time, at ω_r = 1100 rpm (0.5 pu) for different load torques. The slip gain was intentionally set to an incorrect initial value, and the system was brought to a steady state condition at t=0, when the tuning algorithm was enabled. In Fig. 5.9(a), the load torque (T_L) is small, and therefore, the voltage control is dominant. On the other hand, in Fig. 5.9(c) the high T_L results in the dominance of reactive power control. In Fig. 5.9(b) both controls contribute equally for E(pu) computation. Evidently, the fuzzy controller is capable of bringing the system to a tuned condition in approximately 2 sec., in all cases. Similar convergence profiles were observed for other speed conditions, but are not shown here. Fig. 5.10 shows the tuning controller response when the drive is delivering rated torque at ω_r = 0.5 pu, and the rotor resistance is suddenly increased by 100% from the initially tuned condition. Actual tuning time will be much shorter, because detuning range is hardly so large. Since large thermal time constant of the machine causes slow changes of rotor resistance, this tuning time delay is perfectly acceptable.

5.4 Hardware Design

The 5 hp vector control system described in chapter 4 was adapted for the slip gain tuning experimental study, but used with the 5 hp induction servomotor, whose parameters were given on Table 5.3. The major change concerns the machine voltage

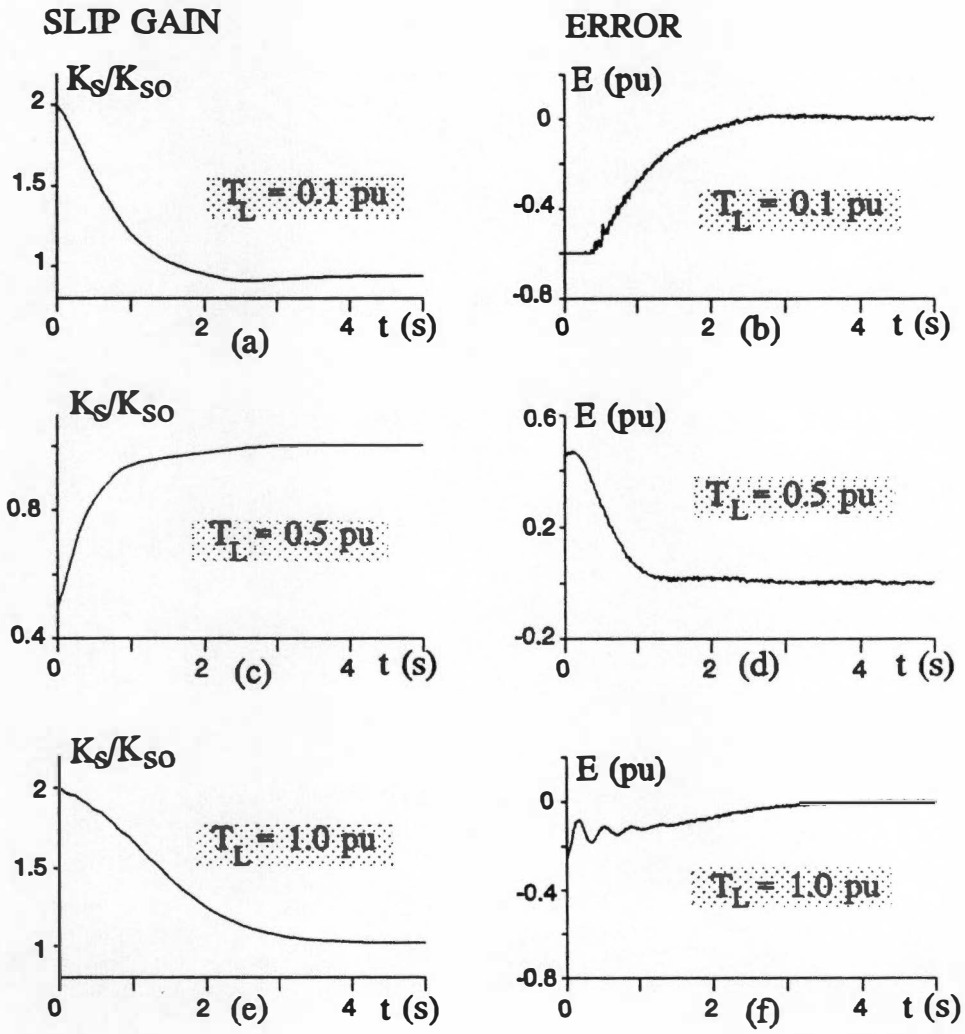


Fig. 5.9 System tuning performance at ω_r=0.5 pu and different torques.

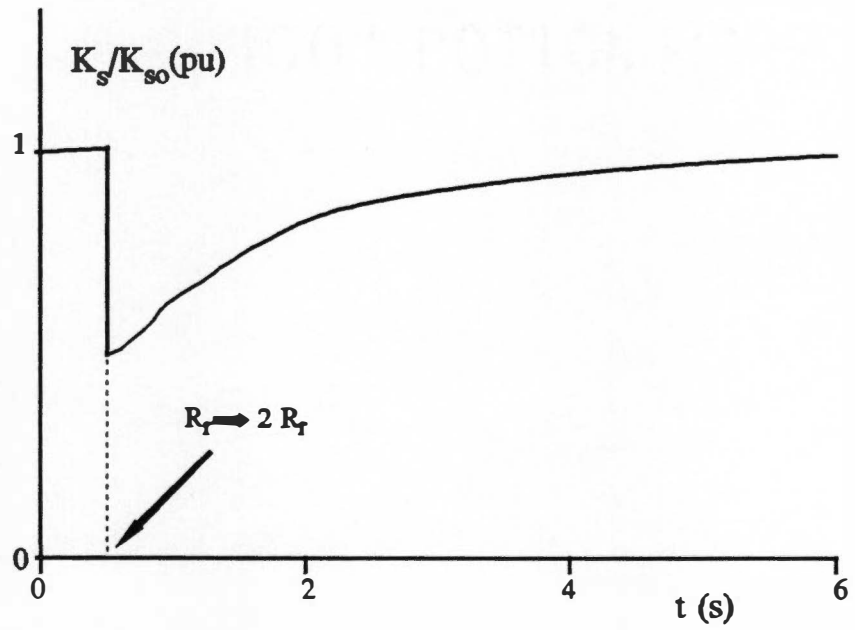


Fig. 5.10 Performance of tuning controller to a 100% increase in rotor resistance.

sensing interface, shown in Fig. 5.11. Its principal component is a 5 KHz bandwidth isolation amplifier (AD 204) from Analog Devices, Inc., that provides complete separation of power and control circuits. The line voltages v_{ac} and v_{cb} are derived by using the AD204 built in operational amplifier in a differential configuration. The large attenuation factor (80:1) of the differential amplifier brings the line voltage signal within the device's input voltage range of ± 5 V. The secondary v_{ac} and v_{cb} signals are next used to compute the stationary frame D^s - Q^s voltages v_{ds}^s and v_{qs}^s, as shown in the following development.

By definition, the D^s - Q^s voltages are obtained from the abc voltages as

$$v_{qs}^s = \frac{2}{3}v_{an} - \frac{1}{3}v_{bn} - \frac{1}{3}v_{cn} \tag{5.19}$$

$$v_{ds}^s = -\frac{v_{bn}}{3} + \frac{v_{cn}}{3} \tag{5.20}$$

Adding and subtracting v_{cn} to (5.19) and rearranging, eqn. 5.21 is obtained

$$v_{qs}^s = (2v_{ac} + v_{cb})/3 \tag{5.21}$$

Rearranging (5.20) directly yields eqn. 5.22

$$v_{ds}^s = v_{cb}/3 \tag{5.22}$$

By using (5.21) and (5.22) in place of (5.19) and (5.20), the neutral line was not required, and only two isolation amplifiers were needed. Furthermore, by performing the abc to D^s-Q^s transformation in hardware, only two A/D channels were needed. Low pass filters were used both at the power and signal sides, to attenuate the PWM related harmonics, as well as to prevent aliasing in the sampling process.

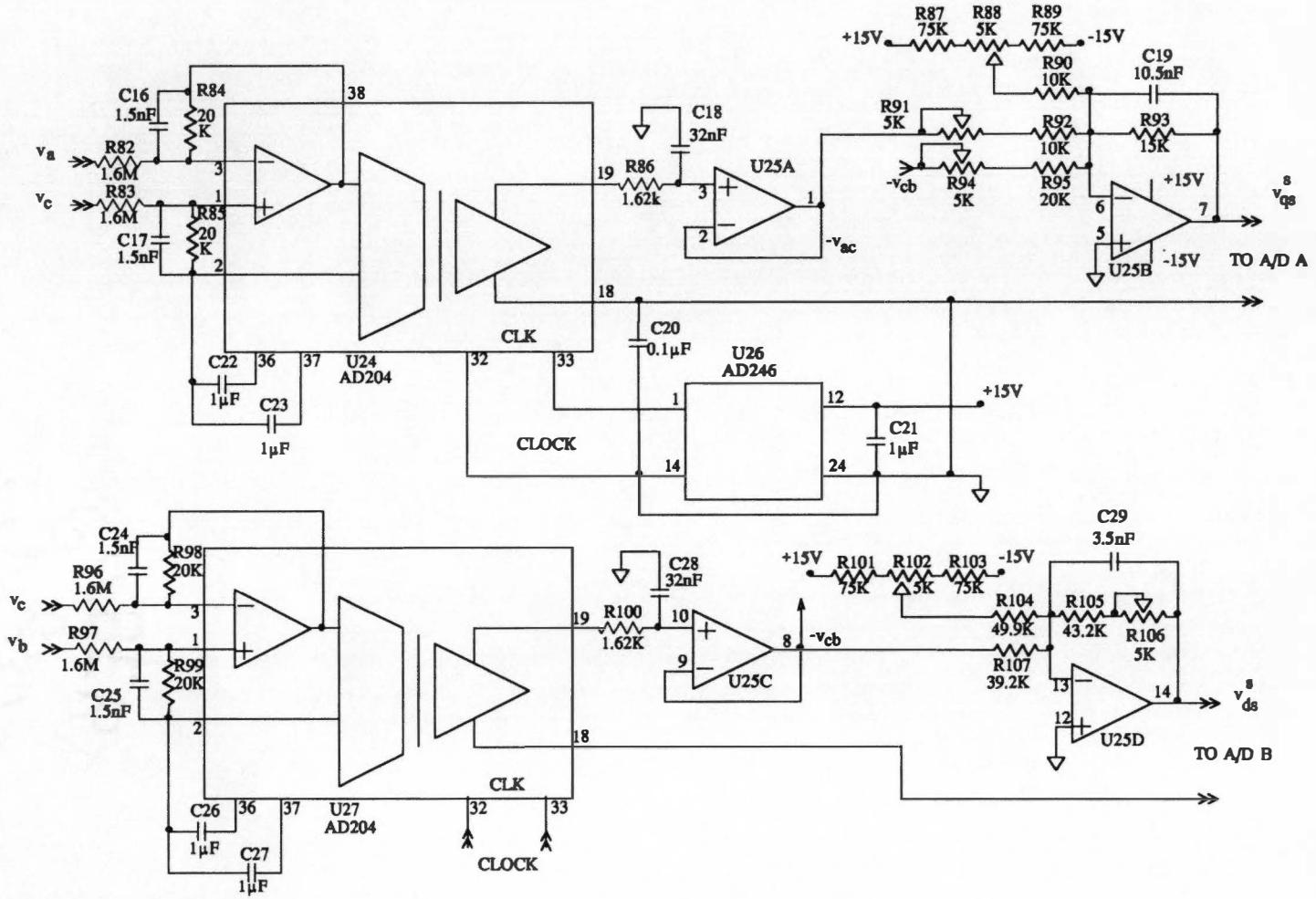


Fig. 5.11 Voltage sensing interface.

The sensing of machine current was not required, since the intrinsic operation of the HBPWM ensures that the actual currents closely track the command values. Therefore, the command currents were used also in the computation of actual v_{ds} and Q quantities.

5.5 Real Time Software Design Issues

The structure of the real time software related to the slip gain tuning is shown in Fig. 5.12. As the vector control routines are the same as those for the efficiency optimization study, they are not shown in the figure.

Proper operation of the tuning controller requires accurate measurement of the machine fundamental voltage. This task is complicated by the variable switching frequency, intrinsic to hysteresis band PWM control operation. This fact led to a conservative design of the analog low pass filters, required by the limited sampling frequency of the DSP. For the particular experimental system used, a cut-off frequency of approximately 1 KHz was selected, that has a negligible impact on the amplitude of the fundamental component, but introduces some amount of phase lag. Without any correction, the phase lag would translate into incorrect values for the synchronous frame voltages, particularly for the v_{ds} signal, that usually has small amplitude. This is turn would cause improper operation of the tuning controller.

A simple steady-state compensation scheme was derived for this problem. For a first order analog low-pass filter of time constant τ, the transfer function is given by

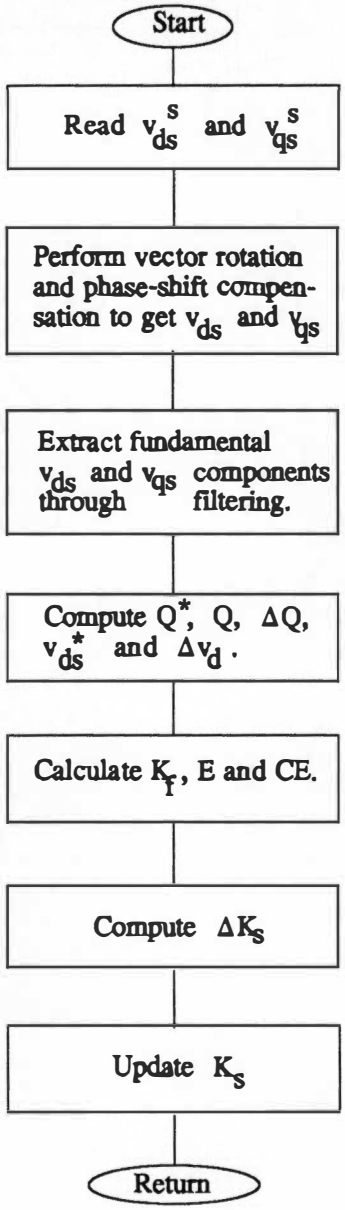


Fig. 5.12 Real time software flowchart.

$$G(s) = \frac{1}{1 + \tau s} \tag{5.23}$$

The phase shift θ_{ps} introduced at frequency $s=j\omega_e$ is determined by

$$\theta_{ps} = \tan^{-1}(\omega_e \tau) \tag{5.24}$$

For small θ_{ps}, $\tan \theta_{ps} \approx \theta_{ps}$, and therefore, $\theta_{ps} \approx \omega_e \tau$. For the case of higher order filters, the same principle can be applied, by computing the equivalent time constant τ_{eq} ($=\Sigma \tau_i$), as long as the overall phase shift remains small. The mechanism of phase shift compensation is illustrated on Fig. 5.13, where only the v_{qs}^s component of stator voltage is considered, for simplicity. As indicated, the measured $\hat{v}_{qs}^s$ lags the correct value by θ_{ps}. By subtracting the estimated θ_{ps} from the original angle θ_e, the q^e-axis is also shifted backwards, and correct phase relationship between $\hat{v}_{qs}^s$ and q^e-axis is restored. The v_{ds} and v_{qs} equations then become

$$v_{ds} = v_{qs}^s \sin(\theta_e - \theta_{ps}) + v_{ds}^s \cos(\theta_e - \theta_{ps}) \tag{5.25}$$

$$v_{qs} = v_{qs}^s \cos(\theta_e - \theta_{ps}) - v_{ds}^s \cos(\theta_e - \theta_{ps}) \tag{5.26}$$

The parameter variation problem in the reference models given in (5.15) and (5.16) is important, since incorrect reference models would try to establish wrong slip gain $\hat{K}_s$. The parameter L_s is susceptible to magnetic saturation, as mentioned before. Even at constant flux command ($i_{ds}^* = i_{dsr}$), use of incorrect slip gain would result in overfluxing

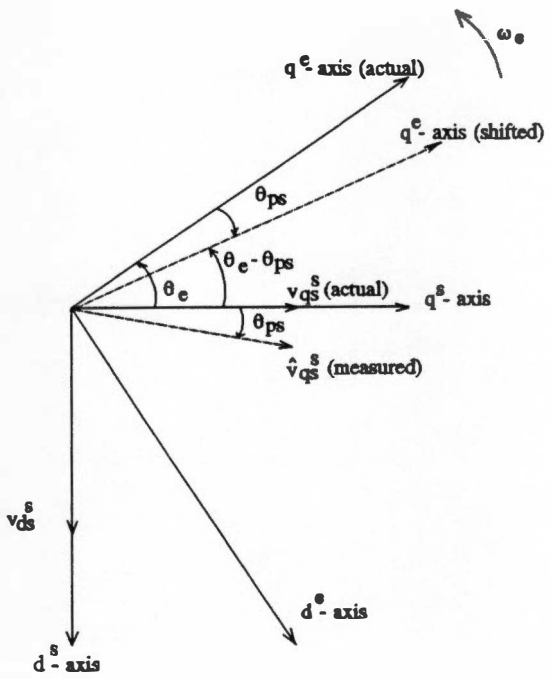


Fig. 5.13 Phase-shift compensation phasor diagram.

(for $\hat{K}_s < K_{so}$) or underfluxing (for $\hat{K}_s > K_{so}$), causing variation on L_s. However, as long as the tuning algorithm possess sufficiently high sensitivity to detuning (as in the proposed method), the error polarity for a large detuning will be coherent, and the controller will drive the estimate $\hat{K}_s$ towards K_{so}, and consequently, force machine flux towards its correct value. This translates into a self correcting effect for L_s, i.e., it will automatically approach its rated value, used in the reference model. It was verified during initial experimentation that the inductance L_σ exhibited significant variation with load torque. To enhance the performance of the controller, this saturation effect was modeled by a quadratic function of the stator current command I_s^*, obtained by least square estimation of the measured saturation profile, as shown in Fig. 5.14. The stator resistance R_s essentially varies linearly with stator temperature, and can be corrected by measurement or estimation of stator temperature. However, no compensation has been provided for this parameter.

5.6 Experimental Study

Experimental performance of the fuzzy tuning controller was thoroughly investigated at various torques and speeds. Prior to the actual tests, the correct slip gain value K_{so} was determined by off-line tests. The controller performance was primarily evaluated by intentionally initializing the slip gain estimate ($\hat{K}_s$) to an incorrect value, activating the tuning controller, and then observing the time domain $\hat{K}_s$ and Error (E) responses. Fig. 5.15 shows the normalized slip gain ($\hat{K}_s/\hat{K}_{so}$) and error (E) responses for $T_L = 0.1$ pu, when $\hat{K}_s$ is initialized to twice the correct value K_{so}, and the tuning algorithm

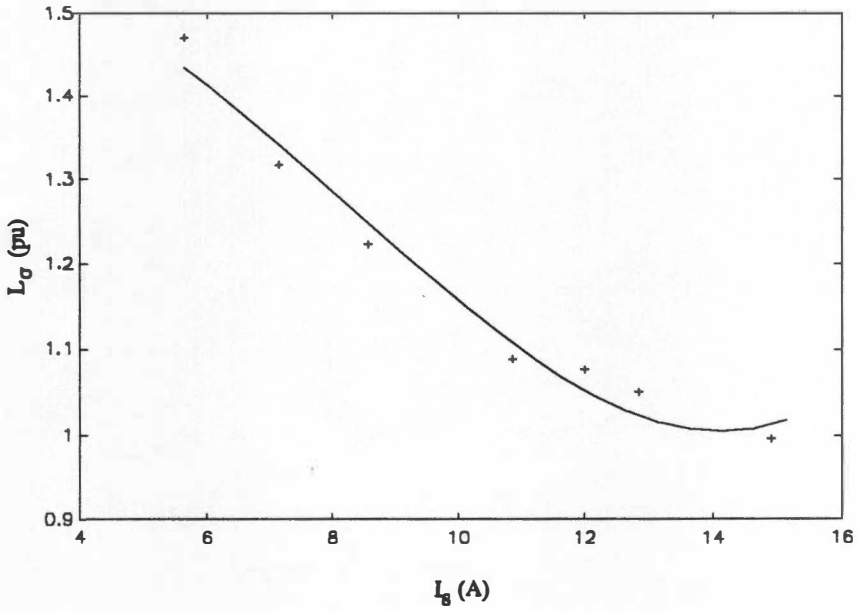
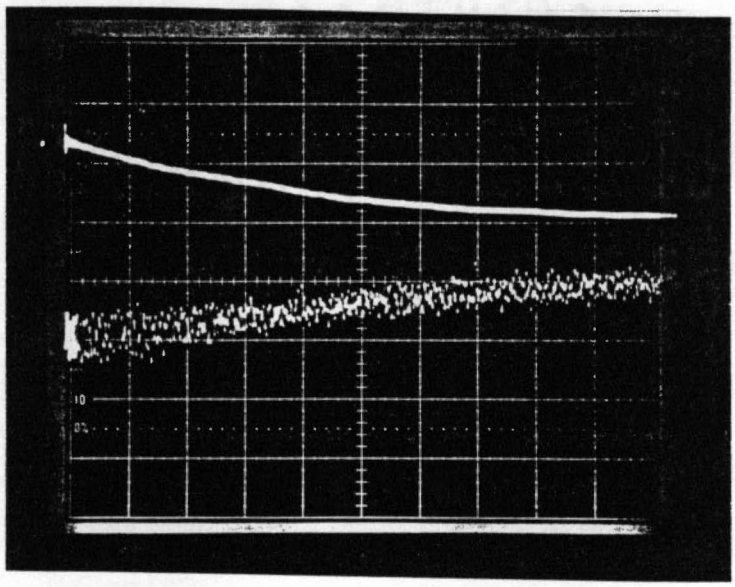


Fig. 5.14 Saturation profile for L_σ.

(a)



(b)

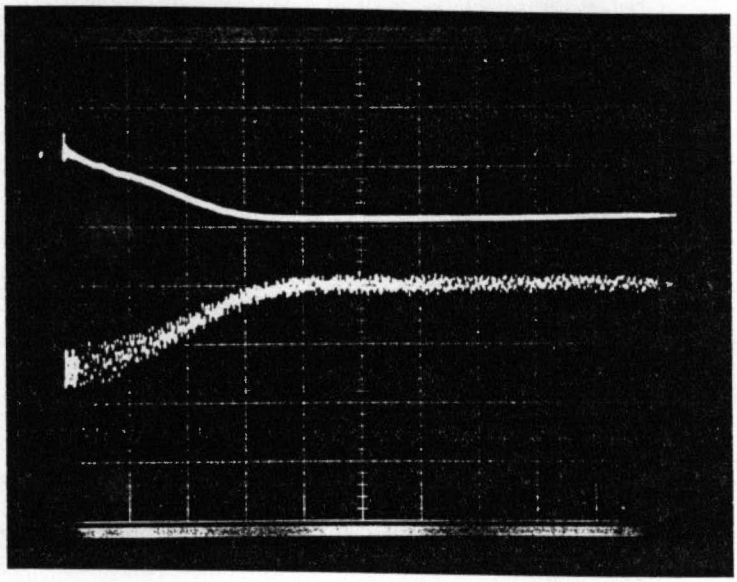


Fig. 5.15 Experimental tuning performance at T_L=0.1 pu and $K_s/K_{so}(0)$=2.
(a) ω_r=0.25 pu. (b) ω_r=0.50 pu.
Top trace: K_s/K_{so}, 0.84 /div.
Bottom trace: Error, 0.2 /div.
Time: 1 sec./div.

is enabled at t=0. In Fig. 5.15(a), ω_r= 0.25 pu, whereas in Fig. 5.15(b) ω_r= 0.5 pu. Fig. 5.16 shows similar responses for T_L= 0.5 pu, when $\hat{K}_s$ is initialized to half K_{so}, and converges to a steady-state value in about 2 sec., with an steady-state error smaller than 9.4% in both cases. Further investigation of proper tuning is shown by the no load speed response to a square wave torque command, present in Fig. 5.17. The waveform of Fig. 5.17(a) reflects a tuned condition ($\hat{K}_s$=$\hat{K}_{so}$), and it differs slightly from the ideal triangular shape, due to the dragging effect of bearing friction and windage. Fig. 5.17(b) shows similar response to a detuned condition ($\hat{K}_s$= 0.5$\hat{K}_{so}$), where the higher order torque dynamics results in speed response degradation, manifested by waveform distortion and amplitude reduction.

In conclusion, a fuzzy logic based tuning method was proposed, that combines the features of MRAC reactive power and d-axis voltage formulations. Fuzzy logic is initially used to determine the dominant method for a given speed and load torque, to ensure a high sensitivity to detuning for the entire torque-speed plane. Next, a second fuzzy controller was used to derive the correcting term ΔK_s from the combined error and error change, that results in a fast convergence under various operating conditions. Both simulation and experimental studies were carried out, demonstrating the effectiveness of the method. The algorithm was able to converge to a good estimate of K_s, with only a small dependency on load torque. However, as the performance of indirect vector control drive systems is insensitive to small slip gain errors [55], this torque dependency does not constitute a practical problem for the application of the proposed technique to actual drive systems.

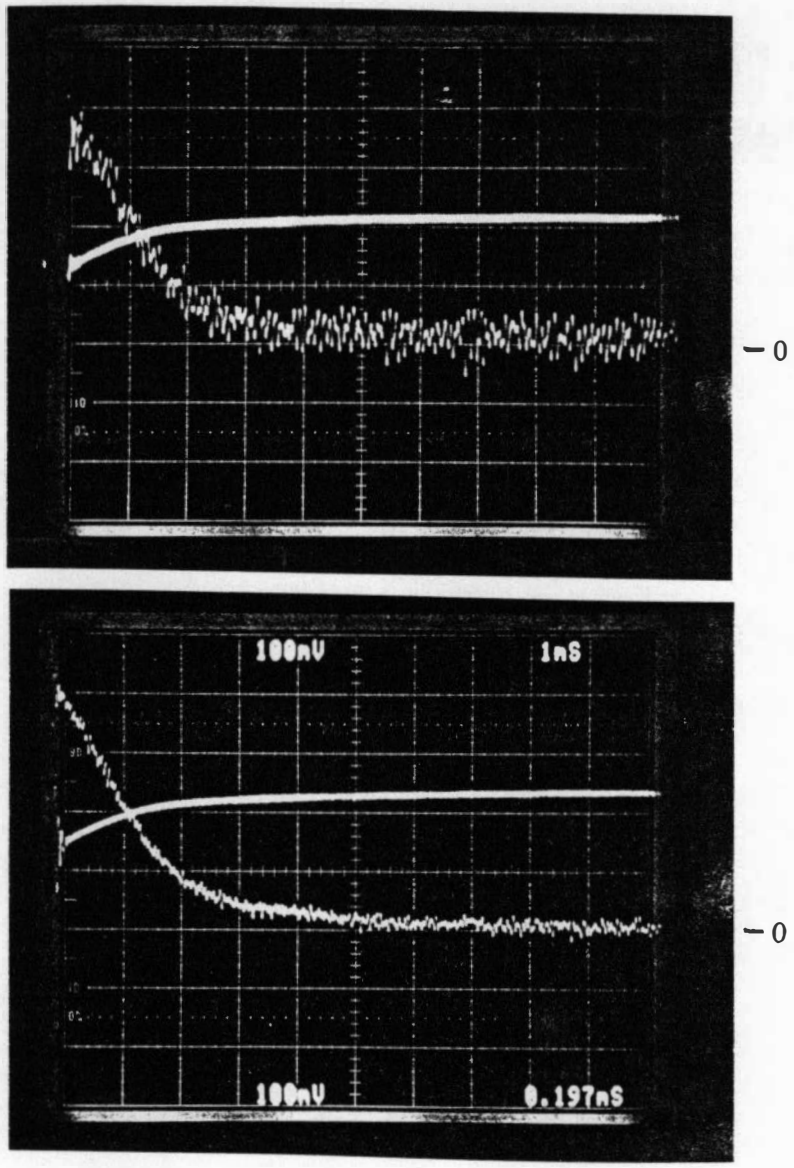


Fig. 5.16 Experimental tuning performance at T_L=0.5 pu and $K_s/K_{so}(0)$=0.5.
(a) ω_r=0.25 pu. (b) ω_r=0.50 pu.
Top trace: K_s/K_{so}, 0.42 /div.
Bottom trace: Error, 0.1 /div.
Time: 0.5 sec./div.

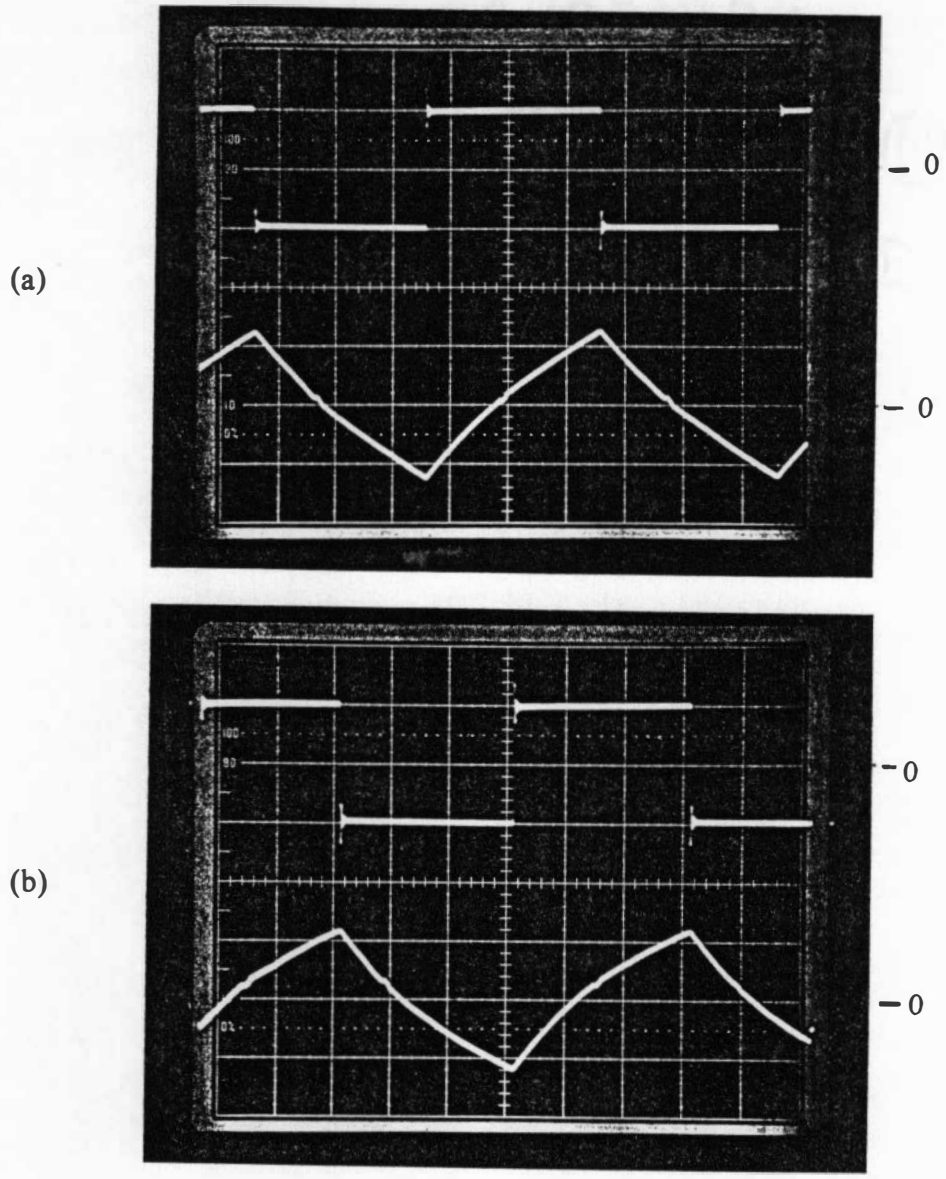


Fig. 5.17 Speed response at no load for a square-wave torque command (i_{qs}^{*}).
(a) K_s/K_{so}=1.0. (b) K_s/K_{so}=0.5.
Top trace: i_{qs}^{*}, 5 A/div.
Bottom trace: ω_r, 286 rpm/div.
Time:100 msec./div.

CHAPTER 6

CONCLUSIONS AND RECOMMENDATIONS FOR FUTURE RESEARCH

The systematic application of fuzzy logic in the control of both DC and AC drives has been discussed in this dissertation. Several control problems have been addressed, using a methodology that typically involved an initial system analysis, followed by control design and validation through digital simulation. Experimental studies were also conducted in most cases, with results that confirm the feasibility of the proposed schemes.

Fuzzy logic control was initially applied to a phase-controlled converter DC drive system that uses a separately excited DC machine. The contributions for this control system are summarized below:

- Study of converter transfer characteristics at continuous and discontinuous conduction modes;
- Development of a fuzzy logic compensation algorithm for the converter system, capable of linearizing converter transfer characteristics during discontinuous conduction mode;
- Design of fuzzy logic controllers for current and speed loops;
- Development of a methodology for simulation of fuzzy controllers in PC-

SIMNON, used in the simulation of fuzzy controllers for both DC and AC drives;

. SIMNON programs design for phase- controlled converter and DC machine;

. Comparative simulation study of fuzzy controlled DC drives and those with conventional PI controls.

The application of fuzzy $\Delta\alpha$ compensation significantly improves the speed of current loop response, particularly when combined with fuzzy current control. Use of fuzzy speed and current controllers in place of conventional PI controllers resulted in a more robust system.

The remaining research effort was devoted to indirect vector controlled induction motor drives. The on-line efficiency optimization via flux control was first considered. In order to validate the control strategy and predict the system performance, a detailed loss model of the converter induction machine system was derived.

The achievements on this subject were:

. Analysis and modeling of induction machine copper losses, core losses and stray losses, under PWM inverter supply;

. Inclusion of skin effect on rotor resistance, temperature effects on winding resistances and modeling of saturation for magnetizing inductance;

. Development of a novel synchronous frame D^e-Q^e lossy equivalent circuit for the induction machine, capable of representing both the loss phenomenon and its dynamic behavior;

. Modeling of conduction and switching losses for a converter system, consisting of diode bridge rectifier and PWM transistor inverter;

. Implementation and testing of machine and converter models in PC-SIMNON, for a 10 hp drive system.

A consistent performance was observed both at steady-state and dynamic conditions. Actual accuracy of lossy models is dependent on the precision with which the machine and converter parameters can be obtained. Currently, one disadvantage of the lossy models is the long simulation time, when using PC-SIMNON. This is especially true when a long term trend is important, such as efficiency optimization studies. Of course, this problem can be minimized by using more powerful computers.

The efficiency optimization study resulted in the following developments:

. Design of a fuzzy logic controller that adaptively adjusts the magnetizing current on the basis of measured DC link power, until the optimum efficiency is achieved;

. Introduction of a feedforward torque compensator to suppress the low frequency pulsating torque, due to changes in flux;

. Development of a criterion for transition between efficiency optimization and transient optimization modes;

. Comprehensive simulation study of a 5 hp drive system with proposed efficiency and torque compensation controls;

. Construction of hysteresis band current controller, protection and monitoring circuits, DC link voltage and current sensing interfaces;

. Design and testing of a speed sensing interface for the TMS320C25 DSP;

. Development of assembly language software for the vector control system;

. Design of assembly language programs for efficiency optimization, torque

compensation and transition controls;

- Experimental evaluation of steady-state and transient performances of proposed control schemes.

The experimental performance correlates well with that predicted by simulation. The efficiency gains are somewhat smaller than expected, because some practical aspects are difficult to model. For example, actual induction machines have some phase unbalance, that creates torque fluctuations and adversely impacts machine efficiency. Irrespective of these problems, the proposed controller ensures true optimum efficiency operation. The method can be incorporated into an existing vector drive system, with a minimum of extra hardware. In most cases, the added software can be easily implemented by the existing processor, resulting in a very cost effective efficiency optimized drive.

The proposed fuzzy efficiency controller is applicable to a number of practical drive systems. In fact, it is currently being considered for an electrical vehicle drive, as part of a technology transfer project to Delco-Remy company.

Finally, fuzzy logic application to the slip gain tuning of an indirect vector-controlled induction motor drive was considered. The salient contributions of the study are:

- An initial fuzzy controller was designed, that combines two MRAC models, D-axis voltage and reactive power, to generate the error signal. This ensures good control sensitivity in the entire torque-speed plane;
- A second fuzzy controller was designed to tune the slip gain based on combined error signal and its slope, resulting in fast convergence at any load condition;

. Simulation programs were developed for the fuzzy tuning controllers and vector control drive, followed by a comprehensive simulation study;

. A hardware voltage interface for v_{ds}^{s} and v_{qs}^{s} sensing was designed and built;

. Assembly language programs were developed for the fuzzy tuning controllers, and experimental validation carried out, for a 5 hp drive system.

The test results correlate well with the theory and simulation performance. All the tuning controls were executed by the same TMS320C25 DSP, used for the basic vector control function. Therefore, the extra cost involved in the fuzzy slip gain tuning controller implementation is reasonably low.

Further research topics to extend the current work would comprehend:

. Use of adjustable DC link voltage topologies to minimize machine harmonic losses;

. Extend the fuzzy tuning control to flux weakening region, along with proper saturation modeling of machine inductances;

. Integration of slip gain tuning technique to efficiency optimization control, to ensure proper pulsating torque compensation as well as fast torque response;

. Investigate the potential use of neural networks to the implementation of fuzzy controllers as well as for better representation of saturation effects.

The fuzzy logic control of drive systems has the potential of enhancing system performance, as well as to reduce sensitivity to disturbances and parameter variation effects. Of course, other control techniques such as Model-Reference Adaptive Control (MRAC), variable structure control (VSC) can also be applied to get robust system

performance. Fuzzy control, however, has the advantages of not requiring a mathematical model of the plant and delivering a chatter free system response.

When compared to other AI techniques, fuzzy control shares some common features with both expert systems and neural networks. While expert systems are ideal for structured knowledge representation and symbolic processing, they possess poor numerical capabilities and are not adequate for real time control of fast power electronic systems. On the other side, neural networks are very efficient in processing unstructured knowledge. Typically, a neural network can be trained to implement a desired mapping, such as a non-linear controller, provided that an extensive numerical input/output data is available to train the network. Although the automated training makes it simpler to tune than a heuristic based fuzzy controller, the training process is not always convergent, and in many cases not very precise. Furthermore, neural networks lack the ability to incorporate qualitative knowledge.

Recently, fuzzy neural networks have been considered. Essentially, neural network techniques are applied in the implementation of parts of a fuzzy controller, such as in the representation of more elaborate membership functions. It can be viewed as an attempt to combine the strengths of both technologies, such as the qualitative knowledge representation of fuzzy logic, with the numerical capabilities of neural networks.

In conclusion, it is expected that, as the research on fuzzy control and other emerging technologies evolves, a formal methodology for fuzzy logic control design and analysis will be developed, along with new CAD tools. Then, the full benefit of this new technology will be realized.

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APPENDICES

APPENDIX A

SIMNON SIMULATION PROGRAMS LISTING

A.1 DC Drive Programs

```
" ------------------------------------------------------------------------
"                                  DC MOTOR
" ------------------------------------------------------------------------
CONTINUOUS SYSTEM DCM

INPUT Vt
OUTPUT W1 Ea  X Ia1 IAPU
TIME T
STATE W Ia IIa
DER DW DIa DIIa

" ---------- Defme state variables.
DW=(-B/J)*W+(Te-TL)/J          "Speed.
DIa=(-Ra/La)*Ia+(Vt-Ea)/La     "Armature current.
DIIa=Ia                        "Integral of Ia, for average computation.
 Te=Kt*Ia                      "Eletromagnetic torque.
TL=KL*W*W                      "Load torque.
Ea=K1*W                        "Counter emf.
Eapu=Ea/Vm

"---------- Auxiliary variables for OUTPUT.
W1=W
X=IIa                          "Used in FCC.
Ia1=Ia                         "Used in RCFR.
Iapu=Ia/53.73

"---------- Parameters.
Kt:0.55                        "Torque constant.
K1:0.55                        "Flux constant.
La:0.008                       "Armature inductance.
Ra:0.6                         "Armature resistance.
J:0.0465                       "Inertia.
B:0.004                        "Viscous damping.
Vm:169.7                       "Base value.
KL:2.78E-04                    "Load constant.
END


" ------------------------------------------------------------------------
"          FUZZY  DELTA ALPHA  COMPENSATION
" ------------------------------------------------------------------------
DISCRETE SYSTEM ACOMP
STATE A  D                     "A, D  Store the previous values of alpha
NEW NA  ND                     " and  delta alpha , respectively.
INPUT ALPHA Ia Iar
```

```
OUTPUT ALP1 ALP2  IDN                "Final values of alpha.

TIME T
TSAMP Z
Z = T +  DT

ALFA=(ALPHA*180)/3.1416             "Alpha in degrees.
IDN=IF Ia>0 THEN Ia/Ibase ELSE 0    "PU current.
J=INT((ALFA+10)/20)                 "Interval indices.
I=INT((IDN+0.01)/0.01)

" ---------- Get degrees of membership and perform rule base evaluation.
MA1=(20*J+10-ALFA)/20
MID1=(0.01*I-IDN)/0.01
MRA=MIN(MID1,MA1)
MRB=MIN(MID1,(1-MA1))
MRC=MIN((1-MID1),MA1)
MRD=MIN((1-MID1),(1-MA1))

" ---------- Compute  value of delta alpha.
IJ=I+J                              "Retrieve contribution of first rule.
DAA=IF I<2 THEN C1 ELSE IF I<3 THEN C2 ELSE IF I<4 THEN C3 ELSE T1
T1=IF I<5 THEN C4 ELSE IF I<6 THEN C5 ELSE IF I<7 THEN C6 ELSE T2
T2=IF I<8 THEN C7 ELSE IF I<9 THEN C8 ELSE IF I<10 THEN C9 ELSE 0
"
C1=IF J<3 THEN 30 ELSE IF J<5 THEN 27 ELSE 25
C2=IF J<2 THEN 5 ELSE IF J<3 THEN 16 ELSE 17
C3=IF J<2 THEN 0 ELSE IF J<3 THEN 10 ELSE 12
C4=IF J<2 THEN 0 ELSE IF J<3 THEN 6 ELSE 9
C5=IF J<2 THEN 0 ELSE IF J<3 THEN 2 ELSE  7
C6=IF J<3 THEN 0 ELSE IF J<4 THEN 4 ELSE 6
C7=IF J<3 THEN 0 ELSE IF J<4 THEN 2 ELSE 4
C8=IF J<4 THEN 0 ELSE 2
C9=IF J<4 THEN 0 ELSE 1
"
L=J+1                               "Retrieve contribution of second rule.
DAB=IF I<2 THEN D1 ELSE IF I<3 THEN D2 ELSE IF I<4 THEN D3 ELSE T6
T6=IF I<5 THEN D4 ELSE IF I<6 THEN D5 ELSE IF I<7 THEN D6 ELSE T7
T7=IF I<8 THEN D7 ELSE IF I<9 THEN D8 ELSE IF I<10 THEN D9 ELSE 0
"
D1=IF L<3 THEN 30 ELSE IF L<5 THEN 27 ELSE 25
D2=IF L<3 THEN 16 ELSE 17
D3=IF L<3 THEN 10 ELSE 12
D4=IF L<3 THEN 6 ELSE 9
D5=IF L<3 THEN 2 ELSE 7
D6=IF L<3 THEN 0 ELSE IF L<4 THEN 4 ELSE 6
D7=IF L<3 THEN 0 ELSE IF L<4 THEN 2 ELSE 4
D8=IF L<4 THEN 0 ELSE 2
D9=IF L<4 THEN 0 ELSE 1
"
M=I+1                               "Retrieve contribution of  third  rule.
DAC=IF M<3 THEN C2 ELSE IF M<4 THEN C3 ELSE IF M<5 THEN C4 ELSE T9
T9=IF M<6 THEN C5 ELSE IF M<7 THEN C6 ELSE IF M<8 THEN C7 ELSE T10
T10=IF M<9 THEN C8 ELSE IF M<10 THEN C9 ELSE 0
DAD=IF M<3 THEN D2 ELSE IF M<4 THEN D3 ELSE T11
T11=IF  M<5 THEN D4 ELSE IF M<6 THEN D5 ELSE T12
T12=IF M<7 THEN D6 ELSE  IF M<8 THEN D7 ELSE T13
T13=IF M<9 THEN D8 ELSE IF M<10 THEN D9 ELSE 0

" ---------- Test of overflow and underflow.
OV=IF J>9 THEN 0 ELSE IF I>9 THEN 0  ELSE 1
UF=IF IJ<2 THEN 0 ELSE 1
```

```
COND=OV*UF*FLAG
SUM=MRA+MRB+MRC+MRD
PROD=DAA*MRA+DAB*MRB+DAC*MRC+DAD*MRD

"---------- Defme final value of alpha depending on sign of E=IAR-IA
ND=DALFA
DK=PROD/SUM
E=Iar-Ia                                "Current loop error.
DALFA=IF COND<1 THEN 0 ELSE IF E>0 THEN CO1 ELSE CO2
CO1=IF DK<D THEN DK ELSE D
CO2=IF DK>D THEN DK ELSE D
ALP1=IF T>0.0028 THEN AU ELSE ALPHA
AU=ALPHA+(DALFA*3.1416)/180             "Alpha in rad.
NA=ALP1
ALP2=A

" ---------- Parameters.
DT:2.778E-03                            "Sampling interval.
Ibase:53.73                             "Base armature current.
FLAG:1                                  "FLAG=1: compensation on;  =0 compensation off.
A:1.5708                                "Initial condition for ALP2.
END


" ------------------------------------------------------------------------------
"                    FUZZY SPEED CONTROLLER
" ------------------------------------------------------------------------------
DISCRETE SYSTEM FSC2

"NOTE:    SEVEN  FUZZY SETS ARE USED FOR STATE VARIABLES E AND CE,
"         WHILE  ELEVEN ARE USED FOR DU.
INPUT W
OUTPUT Iar
STATE X U
NEW NX NU
TIME T
TSAMP TS
TS=T+AT
"
" ---------- Compute error  FE, change in error FCE.
FE=0.10472*WR-W
FCE=FE-X
NX=FE                                   " X = E(k-1)
E=FE/GEs                                "Get the PU values  E and CE.
C=FCE/GCs

" ---------- Compute preliminary values of degree of membership.
A1=MIN(1,(E-Es1)/(1-Es1))
A2=(E-Es2)/(Es1-Es2)
A3=E/Es2
A5=-E/Es2
A6=(-E-Es2)/(Es1-Es2)
A7=(-E-Es1)/(1-Es1)
"
B1=MIN(1,(C-Cs1)/(1-Cs1))
B2=(C-Cs2)/(Cs1-Cs2)
B3=C/Cs2
B5= -C/Cs2
B6=(-C-Cs2)/(Cs1-Cs2)
B7=(-C-Cs1)/(1-Cs1)
```

```
"---------- Get interval indices J and I, for E and CE, respectively.
J=IF E<-ES1 THEN 1 ELSE IF E<-ES2 THEN 2 ELSE IF E<0 THEN 3 ELSE T1
T1=IF E<ES2 THEN 4 ELSE IF E<ES1 THEN 5 ELSE 6
I=IF C<-CS1 THEN 1 ELSE IF C<-CS2 THEN 2 ELSE IF C<0 THEN 3 ELSE T2
T2=IF C<CS2 THEN 4 ELSE IF C<CS1 THEN 5 ELSE 6

" ---------- Compute degree of membership for relevant fuzzy sets.
ME1=IF J<2 THEN MIN(1,A7) ELSE IF J<3 THEN A6 ELSE  T3
T3= IF J<4 THEN A5 ELSE  IF J<5 THEN 1-A3 ELSE IF J<6 THEN 1-A2 ELSE MIN(1,1-A1)
ME2=1-ME1
MC1=IF I<2 THEN MIN(1,B7) ELSE IF I<3 THEN B6 ELSE IF I<4 THEN B5 ELSE T4
T4=IF I<5 THEN 1-B3 ELSE IF I<6 THEN 1-B2 ELSE MIN(1,1-B1)
MC2=1-MC1

" ---------- Perform rule base evaluation.
MRA=MIN(ME1,MC1)
MRB=MIN(ME2,MC1)
MRC=MIN(ME1,MC2)
MRD=MIN(ME2,MC2)

" ---------- Retrieve the control signal for each fired rule.
" ---------- First rule.
D11=IF J<2 THEN C1 ELSE IF J<3 THEN C2 ELSE IF J<4 THEN C3 ELSE T5
T5=IF J<5 THEN C4 ELSE IF J<6 THEN C5 ELSE C6
L=J+1                   "Point to second rule.
D12=IF L<3 THEN C2 ELSE IF L<4 THEN C3 ELSE IF L<5 THEN C4 ELSE T6
T6=IF L<6 THEN C5 ELSE IF L<7 THEN C6 ELSE C7

" ---------- Ci is the i-th  column of the rule base.
C1=IF I<2 THEN -1 ELSE IF I<5 THEN -U1 ELSE IF I<6 THEN -U2 ELSE -U3
C2=IF I<2 THEN -1 ELSE IF I<3 THEN -U1 ELSE IF I<5 THEN -U2 ELSE T6A
T6A=IF I<6 THEN -U3 ELSE 0
C3=IF I<2 THEN -U1 ELSE IF I<3 THEN -U2 ELSE IF I<5 THEN -U3 ELSE T7
T7=IF I<6 THEN 0 ELSE U3
C4=IF I<2 THEN -U1 ELSE IF I<3 THEN -U2 ELSE IF I<4 THEN -U3 ELSE T8
T8=IF I<5 THEN 0 ELSE IF I<6 THEN U3 ELSE U2
C5=IF I<2 THEN -U2 ELSE IF I<3 THEN -U3 ELSE IF I<4 THEN 0 ELSE T9
T9=IF I<6 THEN U3 ELSE U2
C6=IF I<2 THEN -U3 ELSE IF I<3 THEN 0 ELSE IF I<4 THEN U3 ELSE U2
C7=IF I<2 THEN 0 ELSE IF I<3 THEN U3 ELSE IF I<4 THEN U2 ELSE U1
"
M=I+1                                   "Access third and fourth  rules.
D21=IF J<2 THEN VI  ELSE IF J<3 THEN V2 ELSE IF J<4 THEN V3 ELSE T10
T10=IF J<5 THEN V4 ELSE IF J<6 THEN V5 ELSE V6

D22=IF L<3 THEN V2 ELSE IF L<4 THEN V3 ELSE IF L<5 THEN V4 ELSE T11
T11=IF L<6 THEN V5 ELSE IF L<7 THEN V6 ELSE V7

" ---------- Vi also represent the i-th column of the rule base.
V1=IF M<5 THEN -U1 ELSE IF M<6 THEN -U2 ELSE IF M<7 THEN -U3 ELSE 0
V2=IF M<2 THEN -1 ELSE IF M<3 THEN -U1 ELSE T12
T12= IF M<5 THEN -U2 ELSE IF M<6 THEN -U3 ELSE T12B
T12B= IF M<7 THEN 0 ELSE U3
V3=IF M<3 THEN -U2 ELSE IF M<5 THEN -U3 ELSE IF M<6 THEN 0 ELSE T13
T13=IF M<7 THEN U3 ELSE U2
V4=IF M<3 THEN -U2 ELSE IF M<4 THEN -U3 ELSE IF M<5 THEN 0 ELSE T14
T14=IF M<6 THEN U3 ELSE IF M<7 THEN U2 ELSE U1
V5=IF M<3 THEN -U3 ELSE IF M<4 THEN 0 ELSE IF M<6 THEN U3 ELSE U2
V6=IF M<3 THEN 0 ELSE IF M<4 THEN U3 ELSE IF M<7 THEN U2 ELSE 1
V7=IF M<3 THEN U3 ELSE IF M<4 THEN U2 ELSE IF M<7 THEN U1 ELSE 1

" ---------- Defuzzification using height method.
```

```
SUM=MRA+MRB+MRC+MRD
PROD=D11*MRA+D12*MRB+D21*MRC+D22*MRD
DU=PROD/SUM

" ------------------- Compute the next control signal.
UK=U+DU*GUs
IAR=IF UK<UMIN THEN UMIN ELSE MIN(UK,UMAX)
NU=IAR

Wr=IF T<0.8 THEN WR1 ELSE WR2
WR1:1000
WR2:1000

" ------------------- Parameters for the fuzzy controller.
Cs1:0.5
Cs2:0.2
Es1:0.5
Es2:0.2
U1:0.5
U2:0.125
U3:0.05
U4:0.01
UMAX:25
Umin:0
GEs:20
GCs:0.9
GUs:12.5
AT:2.778E-03
END



" ------------------------------------------------------------------------------
"                    FUZZY CURRENT CONTROLLER
" ------------------------------------------------------------------------------
DISCRETE SYSTEM FCC

INPUT Iar Ila Ea
OUTPUT ALPHA1 ALPHA2 Ia Vt
STATE X  U  I   A
NEW NX NU NI NA
TIME T
TSAMP TS
TS=T+AT

NI=Ila
Ia=(Ila-I)/AT                          "Compute average Ia.

" ------------------- Compute error (FE) and change in error (FCE).
FE=Iar-Ia
FCE=FE-X
NX=FE                                  "X = FE(k-1).
E=FE/GEc                               "PU values.
C=FCE/GCc

" ------------------- Compute preliminary degrees of membership.
A1=MIN(1,(E-Ec1)/(1-Ec1))
A2=(E-Ec2)/(Ec1-Ec2)
A3=E/Ec2
A5=-E/Ec2
A6=(-E-Ec2)/(Ec1-Ec2)
A7=(-E-Ec1)/(1-Ec1)
```

```
"
B1=MIN(1,(C-Cc1)/(1-Cc1))
B2=(C-Cc2)/(Cc1-Cc2)
B3=C/Cc2
B5= -C/Cc2
B6=(-C-Cc2)/(Cc1-Cc2)
B7=(-C-Cc1)/(1-Cc1)

" ----------Get interval indices J and H , for E and C, respectively.
J=IF E<-Ec1 THEN 1 ELSE IF E<-Ec2 THEN 2 ELSE IF E<0 THEN 3 ELSE T1
T1=IF E<Ec2 THEN 4 ELSE IF E<Ec1 THEN 5 ELSE 6
H=IF C<-Cc1 THEN 1 ELSE IF C<-Cc2 THEN 2 ELSE IF C<0 THEN 3 ELSE T2
T2=IF C<Cc2 THEN 4 ELSE IF C<Cc1 THEN 5 ELSE 6

" ----------Compute degree of membership for relevant fuzzy sets.
ME1=IF J<2 THEN MIN(1,A7) ELSE IF J<3 THEN A6 ELSE T3
T3= IF J<4 THEN A5 ELSE IF J<5 THEN 1-A3 ELSE T3B
T3B=IF J<6 THEN 1-A2 ELSE MIN(1,1-A1)
ME2=1-ME1
MC1=IF H<2 THEN MIN(1,B7) ELSE IF H<3 THEN B6 ELSE T4
T4= IF H<4 THEN B5 ELSE IF H<5 THEN 1-B3 ELSE T4B
T4B= IF H<6 THEN 1-B2 ELSE MIN(1,1-B1)
MC2=1-MC1

" ---------- Perform rule base evaluation.
MRA=MIN(ME1,MC1)
MRB=MIN(ME2,MC1)
MRC=MIN(ME1,MC2)
MRD=MIN(ME2,MC2)

" ---------- Retrieve the control signal for each fired rule.
" ---------- First rule: (J,H) =(E,C)
D11=IF J<2 THEN C1 ELSE IF J<3 THEN C2 ELSE IF J<4 THEN C3 ELSE T5
T5=IF J<5 THEN C4 ELSE IF J<6 THEN C5 ELSE C6
L=J+1                              "Access second rule: (L,H)=(E,C)
D12=IF L<3 THEN C2 ELSE IF L<4 THEN C3 ELSE IF L<5 THEN C4 ELSE T6
T6=IF L<6 THEN C5 ELSE IF L<7 THEN C6 ELSE C7

" ---------- Ci is the i-th column of rule base.
C1=IF H<2 THEN -1 ELSE IF H<5 THEN -U1 ELSE IF H<6 THEN -U2 ELSE -U3
C2=IF H<2 THEN -1 ELSE IF H<3 THEN -U1 ELSE IF H<5 THEN -U2 ELSE T6A
T6A=IF H<6 THEN -U3 ELSE 0
C3=IF H<2 THEN -U1 ELSE IF H<3 THEN -U2 ELSE IF H<5 THEN -U3 ELSE T7
T7=IF H<6 THEN 0 ELSE U3
C4=IF H<2 THEN -U1 ELSE IF H<3 THEN -U2 ELSE IF H<4 THEN -U3 ELSE T8
T8=IF H<5 THEN 0 ELSE IF H<6 THEN U3 ELSE U2
C5=IF H<2 THEN -U2 ELSE IF H<3 THEN -U3 ELSE IF H<4 THEN 0 ELSE T9
T9=IF H<6 THEN U3 ELSE U2
C6=IF H<2 THEN -U3 ELSE IF H<3 THEN 0 ELSE IF H<4 THEN U3 ELSE U2
C7=IF H<2 THEN 0 ELSE IF H<3 THEN U3 ELSE IF H<4 THEN U2 ELSE U1
"
M=H+1                              "Access third and fourth rules
D21=IF J<2 THEN V1 ELSE IF J<3 THEN V2 ELSE IF J<4 THEN V3 ELSE T10
T10=IF J<5 THEN V4 ELSE IF J<6 THEN V5 ELSE V6
"
D22=IF L<3 THEN V2 ELSE IF L<4 THEN V3 ELSE IF L<5 THEN V4 ELSE T11
T11=IF L<6 THEN V5 ELSE IF L<7 THEN V6 ELSE V7

" ---------- Vi also represents the i-th column of the rule base.
V1=IF M<5 THEN -U1 ELSE IF M<6 THEN -U2 ELSE IF M<7 THEN -U3 ELSE 0
V2=IF M<2 THEN -1 ELSE IF M<3 THEN -UI ELSE  T12
T12=IF M<5 THEN -U2 ELSE IF M<6 THEN -U3 ELSE IF M<7 THEN 0 ELSE U3
```

```
V3=IF M<3 THEN -U2 ELSE IF M<5 THEN -U3 ELSE IF M<6 THEN 0 ELSE T13
T13=IF M<7 THEN U3 ELSE U2
V4=IF M<3 THEN -U2 ELSE IF M<4 THEN -U3 ELSE IF M<5 THEN 0 ELSE T14
T14=IF M<6 THEN U3 ELSE IF M<7 THEN U2 ELSE U1
V5=IF M<3 THEN -U3 ELSE IF M<4 THEN 0 ELSE IF M<6 THEN U3 ELSE U2
V6=IF M<3 THEN 0 ELSE IF M<4 THEN U3 ELSE IF M<7 THEN U2 ELSE 1
V7=IF M<3 THEN U3 ELSE IF M<4 THEN U2 ELSE IF M<7 THEN U1 ELSE 1
"
" ---------- Defuzzification using height method.
SUM=MRA+MRB+MRC+MRD
PROD=D11*MRA+D12*MRB+D21*MRC+D22*MRD
DU=PROD/SUM

" ----------Compute the next control signal.
UK=DU*GUc+U
UKE=UK+Ea
Vt=IF UKE<Umin THEN Umin ELSE MIN(UKE,Umax)
NU=UK+Vt-UKE

" ---------- Calculate firing angle by cosine crossing method.
ALPHA1=ARCCOS(Vt/(1.35*VL))
NA=ALPHA1
ALPHA2=A

" ----------Parameters for the fuzzy controller.
UMAX:110                        "Max. and min. control output.
Umin:-81
AT:2.778E-03                    "Sampling interval.
VL:90                           "Supply line voltage.
Cc1:0.5                         "Change in error fuzzy set definitions.
Cc2:0.2
Ec1:0.5                         "Error fuzzy set definitions.
Ec2:0.2
U1:0.5                          "Control output  fuzzy set definitions.
U2:0.125
U3:0.05
"
GEc:25                          "Input and output gains.
GCc:5
GUc:40

"----------Initial values
A:1.5708
END



" ------------------------------------------------------------------------
"              PHASE-CONTROLLED RECTIFIER
" ------------------------------------------------------------------------
CONTINUOUS SYSTEM RCFR
"
INPUT ALPHA1 ALPHA2 Ia Ea
OUTPUT Vt IVT VTPU
STATE IV                        "USED TO GENERATE Vdc IN CPI
DER   DIV
TIME T
TSH=T-N*2.778E-03               "Relative time.
N=INT((T-1E-10)/2.778E-03)

"-----DEFINE BASIC VOLTAGES PROFILES.
Vab=Vmax*SIN(1.0472+W*TSH)
```

```
Vcb=Vmax*SIN(2.0944+W*TSH)
Vca=Vmax*SIN(3.1416+W*TSH)
"
" ----------FIRING ANGLE REFLECTED IN THE INTERVAL 0 - 60 DEG.
UA=IF ALPHA1<1.0472 THEN ALPHA1 ELSE ALPHA1-1.0472
FT=UA/W                                    "FIRING TIME
Vold= IF ALPHA1<1.0472 THEN Vcb ELSE Vca
Vnew= IF ALPHA1<1.0472 THEN Vab ELSE Vcb
VT1=IF TSH<FT THEN Vold ELSE Vnew
VT=IF VT1>Ea THEN VT1 ELSE CON1            "Converter output voltage.
CON1=IF Ia>1E-03 THEN VT1 ELSE Ea          "Test for continuous conduction.

" ----------EVALUATE INTEGRAL OF VT
Vtpu=Vt /Em
DIV=Vtpu                                   "Vt (pu) integral.
IVT=IV                                     "Electrical frequency.

" ----------Parameters.
W=376.99
Vmax:127.28                                "Peak line voltage.
END


" ------------------------------------------------------------------------
"          DISCRETE PI CURRENT CONTROLLER
" ------------------------------------------------------------------------
" Note: INCORPORATES ANTI-WINDUP.
DISCRETE SYSTEM CPI
INPUT Iar IIa I1 IVT Ea                    "I1 is the actual current and IIa is its integral.
OUTPUT ALPHAI ALPHA2 Ia Vt VDC

STATE X   I   Vd   A                       " A = ALPHA(k-1)
NEW NX NI NVd NA
TIME T
TSAMP TS
TS=T+AT

" ----------Computation of Idc and Vdc (pu).
NI=IIa
Ia=(IIa-I)/AT
NVd=IVT
Vdc=(IVT-Vd)/AT

" ----------COMPUTE ERROR E, CONTROL SIGNAL U
E=Iar-Ia
V=Kpi*E+X+Ea
U=IF V<Umin THEN Umin ELSE IF V>UMAX THEN UMAX ELSE V
NX=X+Kii*E*AT
Vt=U

ALPHA1=ARCCOS(Vt/(1.35*VL))                " Firing angle.
NA=ALPHA1
ALPHA2=A                                   " Alpha(k-1)

" ---------- PARAMETERS
UMAX:110                                   "Max. and min. control output.
Umin:-81
Kpi:1.2                                    "PI gains.
Kii:80
AT:2.778E-03                               "Sampling interval.
VL:90                                      "Line voltage.
" ----------INITIAL VALUES.
```

```
A:2.094                          "120 DEGREES
END


" ------------------------------------------------------------------------
"              DISCRETE PI  SPEED  CONTROLLER
" ------------------------------------------------------------------------
"NOTE:  Using anti-windup.
DISCRETE SYSTEM SPI
INPUT W
OUTPUT Iar

STATE X
NEW NX
TIME T
TSAMP TS
TS=T+ADT

" ---------- COMPUTE ERROR E AND CONTROL SIGNAL U.
E=0.10472*Wr-W
V=Kps*E+X
U=IF V<0 THEN 0 ELSE IF V>UMAX THEN UMAX ELSE V
NX=X+Kis*E*ADT+U-V "WITH ANTI WINDUP
Iar=U                                "Reference current.
Wr=IF T<0.8 THEN Wr1 ELSE Wr2        "Reference speed.

"--------PARAMETERS.
UMAX:25                              "Maximum current reference.
Kis:28                               "PI gains.
Kps:2
ADT:2.778E-03                        "Sampling interval.
Wr1:1000
Wr2:1000
END
```

A.2 Loss Modeling and Efficiency Optimization Programs

```
" ------------------------------------------------------------------------
"                        AC POWER SUPPLY
" ------------------------------------------------------------------------
CONTINUOUS SYSTEM AC
OUTPUT vas vbs vcs we
TIME t

"----Generate 3-phase sinusoidal voltages
vas=Vm*sin(wt)
vbs=Vm*sin(wt-2.0944)
vcs=Vm*sin(wt-4.1888)
wt=we*t
we=377
Vm:187.79
END
```

```
" ------------------------------------------------------------------------------------------
"  LOSSY MODEL OF A THREE-PHASE DIODE BRIDGE RECTIFIER
" ------------------------------------------------------------------------------------------
CONTINUOUS SYSTEM RCFL
INPUT  Vas Vbs Vcs I2
OUTPUT Vd Pild  Pior Piir
STATE Vc    I1
DER   DVc DI1
TIME t

"---------------------GENERATION OF IDEAL RECTIFIER VOLTAGE Vr
Vp = MAX(Vas, MAX(Vbs,Vcs))          "Positive profile,
Vn = MIN(Vas, MIN(Vbs,Vcs))          "Negative profile
Vr = Vp-Vn

"---------------------EQUATIONS FOR CURRENT AND VOLTAGES
Vdd = Vd0+Cd*ABS(I1)^Ed              "Diode voltage drop
DVc = (I1-I2)/C                      "Capacitor voltage
DI1 = IF I1<0 AND AUX<0 THEN 0 ELSE AUX  "Current through L
AUX = (Vr-2*Vdd-Vc)/L                "Voltage across L
Vd = Vc

"---------------------INSTANTANEOUS POWERS AND LOSSES
Pild = 2*Vdd*abs(I1)                 "Instantaneous condution losses in diodes
Pior = Vd*I1                         "Instantaneous power output of rectifier
Piir = Vr*I1                         "Instantaneous power input of rectifier

"-------INITIAL VALUES
Vc:314
I1:1.57

"-------PARAMETERS FOR POWEREX CD 411230 (30 A/ 1200-1600 V)
C:0.0048                  "DC link capacitor
L:0.001                   "DC link equivalent of AC source inductance
Cd:0.052                  "Volt-ampere coefficient of diode characteristics
Ed:0.585                  "Exponent of diode characteristics
Vd0:0.8                   "Forward voltage drop of diode
END



" ------------------------------------------------------------------------------------------
"          LOSSY MODEL OF  A THREE-PHASE HYSTERESIS
"                          BAND PWM INVERTER
" ------------------------------------------------------------------------------------------
"
DISCRETE SYSTEM INVL
INPUT Iac Ibc Icc Ial Ibl Icl Vd We AT
OUTPUT Va Vb Vc I2 Nsaux Plinv
STATE A   B   C   sl   s   SIa   Iav          "Inverter Switching State
NEW  NA NB NC Nsl Ns NSIa NIav                "A=1 -> T1 on, T4 off
STATE Ia    Ib    Ic
NEW   NIa  NIb  NIc
TIME T                                        "A=0 -> T4 on, T1 off
TSAMP Ts
Ts = T + AT

"-------DEFINE AUXILIAR STATE EQUATIONS FOR CURRENT
NIa = Ial
NIb = Ibl
NIc = Icl
```

```
"-------DEFINITION OF ERROR = REF-ACTUAL
Iae = Iac-Ia
Ibe = Ibc-Ib
Ice = Icc-Ic

"-------LOGIC TO GENERATE PWM BY HB COMPARISON
NA = IF Iae>HB THEN 1 ELSE IF IAE<-HB THEN 0 ELSE A
NB = IF Ibe>HB THEN 1 ELSE IF IBE<-HB THEN 0 ELSE B
NC = IF Ice>HB THEN 1 ELSE IF ICE<-HB THEN 0 ELSE C

"-------GENERATION OF VOLTAGES (WITH RESPECT TO GROUND)
Vag = IF Ia>0  THEN C1a ELSE C2a
C1a = IF NA>0 THEN Vd-TdropA ELSE -DdropA
C2a = IF NA>0 THEN Vd+DdropA ELSE TdropA
TdropA = ABS(Ia)*Rt + Vt0
DdropA = Vd0 + Cd*ABS(Ia)^Ed

Vbg = IF Ib>0 THEN C1b ELSE C2b
C1b = IF NB>0 THEN Vd-TdropB ELSE -DdropB
C2b = IF NB>0 THEN Vd+DdropB ELSE TdropB
TdropB = ABS(Ib)*Rt + Vt0
DdropB = Vd0 + Cd*ABS(Ib)^Ed

Vcg = IF Ic>0 THEN C1c ELSE C2c
C1c = IF NC>0 THEN Vd-TdropC ELSE -DdropC
C2c = IF NC>0 THEN Vd+DdropC ELSE TdropC
TdropC = ABS(Ic)*Rt + Vt0
DdropC = Vd0 + Cd*ABS(Ic)^Ed

"-------GENERATION OF PHASE VOLTAGES
Va = (2*Vag-Vbg-Vcg)/3
Vb = (2*Vbg-Vag-Vcg)/3
Vc = (2*Vcg-Vbg-Vag)/3

"-------CONDUCTION LOSSES COMPUTATION
Pca = IF NA THEN (Vd-Vag)*Ia ELSE -Vag*Ia
Pcb = IF NB THEN (Vd-Vbg)*Ib ELSE -Vbg*Ib
Pcc = IF NC THEN (Vd-Vcg)*Ic ELSE -Vcg*Ic
Picli = Pca+Pcb+Pcc                    "Instantaneous conduction losses

"-------CONTABILIZATION OF SWITCHING CYCLES/PERIOD
"       FOR UPPER TR OF LEG A. (Nsaux)
RES = SIGN(Iac)                        "LOGICAL SIGNAL (+1 OR -1)
DRES = DELAY(RES,AT)                   "DELAYED LOGICAL SIGNAL
PROD = RES*DRES                        "PULSE FOR RESET (-1 AT RESET INSTANT)
Ns1 = IF PROD>0 THEN (IF A<NA OR A>NA THEN s1+1 ELSE s1) ELSE 0
Ns = IF PROD<0 THEN 2*s1 else s
Nsaux = s

"------COMPUTE AVERAGE VALUE OF ABS(Ia)
fs = We/pi2
NSIa = IF PROD>0 THEN SIa+ ABS(Ia) ELSE 0    "Integral over half period.
NIav = IF PROD>0 THEN Iav ELSE (2*fs*AT*SIa)          "Average over half cycle.

"-------COMPUTATION OF TRs TURN-ON SWITCHING LOSSES
Nif = (Nsaux/2)*Iav*fs
Kon = 1.5*ton*(1-1.333*(t1L/ton)+0.5*(t1L/ton)*(t1L/ton))
Rton = Vd/(Nif*Kon+alpha)
Pton = Vd*Vd/Rton                      "Turn on losses.
Iton = Vd/Rton                         "Current component due to Pton

"-------COMPUTATION OF TRs TURN-OFF SWITCHING LOSSES
```

```
Koff = 1.5*toff*(1-1.333*(t1C/toff)+0.5*(t1C/toff)*(t1C/toff))
Rtoff = Vd/(Nif*Koff+alpha)
Ptoff = Vd*Vd/Rtoff
Itoff = Vd/Rtoff                        "Current component due to Ptoff

"-------COMPUTATION OF SNUBBER LOSSES
Rsn = 2/(3*Nsaux*fs*Cs+alpha)
Psn = Vd*Vd/Rsn
Isn = Vd/Rsn                            "Current component due to Psn

"-------COMPUTATION OF DC LINK CURRENT AND INV. OUTPUT POWER
I2 = NA*Ia+NB*Ib+NC*Ic+ (Iton+Itoff+Isn)
Plinv = Picli + Pton + Ptoff + Psn      "Total inverter losses.

"-------INITIAL VALUES
A:1                         "Base drive states.
B:1
C:0
s1:1
s:1
Iav:10

"-------PARAMETERS
"Data for POWEREX single darlington TR ks524503 (30 A/ 600 V)
toff: 0.8e-06                           "Turn-off time of TR ( tf )
ton: 0.52e-06                           "Turn-on time of TR ( tr )
Rt: 0.01                                "Conduction resistance of TR
Vt0: 0.7                                "Saturated  coletor-emitter of TR
Vd0: 0.7                                "Forward voltage drop of freewheeling diode
Cd: 0.051                               "Coefficient of VxA characteristic of FWD
Ed: 0.671                               "Exponent of VxA  charac. of FWD

"Data related to snubbers
Cs: 0.0068E-06                          "Snubber capacitor
Ls: 0.2e-06                             "Snubber inductance
t1L:0.11e-06
t1C:0.382e-06

"Other parameters
alpha:1e-07
pi2= 8*atan(1)
HB: 1.5                                 "Histeresis band
END



"  ------------------------------------------------------------------------------------------
"     INDUCTION MOTOR LOSSY D-Q MODEL IN SYNCHRONOUS FRAME
"  ------------------------------------------------------------------------------------------
" FEATURES INCLUDED IN THIS MODEL
"  1) TEMPERATURE EFFECTS ON BOTH ROTOR AND STATOR
"       RESISTANCES ;
"  2) SATURATION EFFECTS ON MAGNETIZING INDUCTANCE (Lm);
"  3) SKIN EFFECT ON ROTOR RESISTANCE (HARMONIC FREQUENCY);
"  4) STRAY LOSSES (BOTH FUND. AND HARMONICS) ARE INCLUDED;
"  5) FRICTION AND WINDAGE LOSSES ARE CONSIDERED;
"  6) CORE LOSSES ARE SEPARATELY COMPUTED FOR FUND. AND
"       HARMONIC FREQUENCIES AND INTEGRATED IN THE MODEL.
"
CONTINUOUS SYSTEM IMS3
INPUT  va vb vc we Nsaux SINWT COSWT
OUTPUT ia ib ic wr Lm  Rr po Pin
```

```
STATE iqsl idsl iqsr idsr iqm idm iqrl idrl
DER  diqsl didsl diqsr didsr diqm didm diqrl didrl
STATE iqrr idrr w   Plf  fc
DER   diqrr didrr dw DPlf dfc
TIME t

"-------DEFINE SOME USEFUL CONSTANTS AND VARIABLES
pi = atan(1)*4
p2 = pi*2
pb2s = p2*p2
sq2 = sqrt(2)
sqr3 = sqrt(3)

tl = kl*wr*wr                          "Load torque
spd = wr*60/p2                         "Speed in rpm
wr=w*2/pole                            "In mech. rad./s
f = we/p2                              "Supply frequency
fca = max(f*Nsaux,2500)                "Harmonic frequency (carrier)
wc = p2*fc                             "Harmonic frequency (in rad/s)
dfc = (fca-fc)/talt

"-------COMPUTE MACHINE FLUXES
fqls = Lls*iqsl                        "Stator leakage fluxes
fdls = Lls*idsl
fqm = Lm*iqm                           "Air-gap fluxes
fdm = Lm*idm
fqr = fqm - Llr*iqrl                   "Rotor fluxes
fdr = fdm - Llr*idrl

"-------DEFINE CURRENTS AND AUXILIAR VARIABLES
iqs = iqsl + iqsr
iqr = -(iqrl + iqrr)
iqrm = iqs + iqr - iqm
F1q = -vqs + R1*iqs + we*fdls
F2q = (RbL2*iqsr-F1q*Leq2-RbL3*iqrm-we*fdm*Leq3+wsl*fdr*Leq4+F2qc)/Leq1
F2qc = (RbL5+RbL6)*iqr-RbL4*iqrr
LF2q = Lls*F2q

ids = idsl + idsr
idr = -(idrl + idrr)
idrm = ids + idr - idm
F1d = -vds + R1*ids - we*fqls
F2d = (RbL2*idsr-F1d*Leq2-RbL3*idrm+we*fqm*Leq3-wsl*fqr*Leq4+F2dc)/Leq1
F2dc = (RbL5+RbL6)*idr-RbL4*idrr
LF2d = Lls*F2d

"-------STATE EQUATIONS FOR MACHINE DYNAMICS
diqsl = F2q                            "Q-axis circuit
diqsr = (LF2q - Rssn*iqsr)/Lssn
diqm = (-F1q-LF2q - we*fdm)/Lm
Vlrq = -F1q-LF2q + wsl*fdr-we*fdm+Rr*iqr
diqrl = Vlrq / Llr
diqrr = (Vlrq - Rrsnp*iqrr)/Lrsn

didsl = F2d                            "Q-axis circuit
didsr = (LF2d - Rssn*idsr)/Lssn
didm = (-F1d-LF2d + we*fqm)/Lm
Vlrd = -F1d-LF2d -wsl*fqr + we*fqm+Rr*idr
didrl = Vlrd / Llr
didrr = (Vlrd - Rrsnp*idrr)/Lrsn

dw = (te-tl-b*wr-Kfw*Wr*Wr)*pole/(j*2)  "In elec. rd/s
```

```
"——-VOLTAGE TRANSFORMATIONS
vqss = va                                   "abc to d-q stationary
vdss = (vc-vb)/sqr3
Vqs = vqss*coswt-vdss*sinwt                 "d-q stationary to synchronous
Vds = vqss*sinwt+vdss*coswt

"-------COMPUTATION OF DEVELOPED TORQUE
te = -(fdr*iqr-fqr*idr)*0.75*pole           "Instantaneous value

"-------CURRENT TRASFORMATIONS
iqss = coswt*iqs+sinwt*ids                  "synchonous to stationary
idss = -sinwt*iqs+coswt*ids
ia = iqss                                   "d-q stat. to abc
ib = -(0.5*iqss)-(0.5*sqr3*idss)
ic = -(0.5*iqss)+(0.5*sqr3*idss)

"-------CALCULATE THE RMS CURRENT THROUGH Lm
im = sqrt(iqm*iqm+idm*idm)/sq2

"-------COMPUTE TIME DEPENDENT MODEL PARAMETERS
rs = rs0*(1+A1*Dtemp)                       "Resistance values correct
rr = rr0*(1+A2*Dtemp)                       "for temperature
Lm = if im<im0 then lm0 else lm0-m*(im-im0) "Magnetizing ind.
Lr = Lm+Llr

wsl = we-w                                  "Slip frequency
s =  wsl/we                                 "Per unit slip
Rm = pb2s*f/((1+abs(s))*Kh+(1+s*s)*Ke*f)    "Core loss resistor (fund.)
Rmc = pb2s*fc /(2*(Kh+Ke*fc))               "Core loss resistor (harm.)
Rtemp = Rm*(Rmc-Rm)
Llm = if Rtemp > 0.18 then sqrt(Rm*(Rmc-Rm))/wc else 10e-06
                                            "Leakage mag. induc.(harm.)

" ---------- Parameter for stray loss in stator
Rsll = Rslb*(f/fb)*(1+gm*f)/(1+gm*fb)       "Series resist. (fund.)
Cfr = (fc/fcb)*(1+gm*fcb)/(1+gm*fc)         "Correction factor for harm.
Rssn = Rssb*Cfr                             "Stator harm. stray loss res.
Rrc = Rr*Kr*sqrt(fc)                        "Rotor resist. at harm. freq.

"   Parameters for stray loss and skin effect in the rotor
Xlrc = Llr*wc                               "Rotor leakage reac. at harm.freq.
Xlrc2 = Xlrc*Xlrc                           "Square of Xlrc
Rrsn = Cfr*Rrsb                             "Rotor harm. stray loss res.
Rrsnb = Xlrc2* Rrsn / (Rrsn*Rrsn+Xlrc2)     "Series eq. of Rrsn (harm. freq.)
Rrle = Rrsnb+(Rrc-Rr)                       "Combined stray + skin res.,series
XbR = Xlrc2/Rrle                            "Aux. variable
SIG = If Xlrc < Rrsn then +1 else -1        "Sign of correct root of quadr. expr.
DLT = max((XbR*XbR - 4*Xlrc2),0)            "Aux. variable
Rrsnp = If DLT>0 then (XbR+SIG*sqrt(DLT))/2 else Xlrc/2
                                            "Parallel combined stray+skin res.

"-------COMPUTE OUTPUT AND INPUT POWERS
po = tl*wr                                  "Output power
pin = va*ia+vb*ib+vc*ic                     "Input  power, instantaneous
Pl =  pin-po                                "Instantaneous overall losses

"-------ESTIMATE TEMPERATURE RISE
DPlf = (Pl-Plf)/Talt                        "Average overall losses
Dtemp = Plf*theta

"-------DEFINE AUXILIAR CONSTANTS
Leq4 = 1/Lrsn + 1/Llr
```

```
Leq3 = Leq4 + 1/Lm
Leq2 = Leq3 + 1/Llm
Leql = (Leq2 + 1/Lssn)*Lls + 1
RbL2 = Rssn /Lssn
RbL3 = Rm/Llm
RbL4 = Rrsnp/Lrsn
RbL5 = Rr/Llr
RbL6 = Rr/Lrsn
Rl = Rs + Rsll

"-------CONSTANT MACHINE PARAMETERS
" CLASS B 5 hp machine.
rs0:0.370                       "Resistance values at 25C, 60 Hz
rr0:0.436                       " "       "   " ", dc
Rslb:0.1384                     "Base value of series fund. stray loss stator res.
Rssb:164.46                     "Base value of stator harm. stray loss res. (at fcb=5KHz)
Rrsb:164.46                     "Base value of rotor harm. stray loss res. (at fcb)

theta: 0.06319                  "Thermal resistance
A1:0.00385                      "Temperature coefficient for copper 100
A2:0.00389                      " "       "     for aluminum 60

kh:4.3811                       "Hysteresis coefficient
ke:0.0313                       "Eddy current coefficient
gm:0.00238
Kr:0.1371                       "Skin effect coef. for 3/8" deep bar
fb:60                           "Base frequency (1800 rpm)

lm0:0.06277                     "Linear value of magnetizing inductance
im0:3.4                         "Linear region breakpoint of Lm
m:0.00366                       "Saturation coefficient
Lls:0.00213                     "Leakage inductances
Llr:0.00213
Lssn = 0.05*Lls
Lrsn = 0.05*Llr

j:17.5E-03                      "Machine inertia
b:4.05e-04                      "Viscous damping coefficient
pole:4
kl = factor*0.5862e-03          "Load constant (Rated = 0.5862e-03)
factor = if t<3 then 0.4 else 2 !
Kfw:2.248e-06                   "Friction and windage coefficient
Tal:0.0004                      "Time constant for filters
Talt:0.010
fcb:5000                        "Base carrier frequency

"-------INITIAL CONDITIONS FOR Wref=900 rpm
idsl:9.786
idm: 9.7
iqsl:1.6
iqrl:1.6
w:188.5                         "1800 rpm  --> 377 rad/s.
fc:5000
END
```

```
" ------------------------------------------------------------------------------------------
"       DISCRETE  SPEED PI CONTROLLER
" ------------------------------------------------------------------------------------------
" NOTES:
"          1) ANTI-WINDUP IS USED, AS WELL AS SATURATION FUNCTION
"              TO LIMIT VALUE OF CONTROL VARIABLE iqec;
"          2) GAINS ARE SELECTED FOR A 5 HP MACHINE;

DISCRETE SYSTEM WCTRL
INPUT Wr
OUTPUT E  iqecp
STATE X
NEW NX
TIME T
TSAMP TS
TS=T+AT

"--------COMPUTE SPEED ERROR E AND INITIAL CONTROL SIGNAL V
E = 0.1047197*Wref-Wr
V = Kp*E+X
NX = X+Ki*E*AT+kwup*(iqecp-V)

"--------COMPUTE FINAL CONTROL SIGNAL iqec
iqecp = IF ABS(V)<IQM THEN V ELSE SIGN(V)*IQM

"--------PARAMETERS
Wref:2200
IQM:26.7                          "Maximum iqec  (50% higher Is).
Kp:1.25                           "Proportional gain
Ki:25                             "Integral gain
Kwup:0.02                         "Anti-windup coefficient.
AT:1E-03                          "Speed sampling time.

"--------INITIAL VALUES
X:5
END


" ------------------------------------------------------------------------------------------
"                FUZZY EFFICIENCY OPTIMIZATION ROUTINE
" ------------------------------------------------------------------------------------------
DISCRETE SYSTEM FEOPT
INPUT  DWr Wr Pi DT TLest
OUTPUT DIDEC RT FSSA
STATE  Po  DPi  CR  DC  Dids  FSS        " DC=Down Counter
NEW  NPo NDPi NCR NDC NDids NFSS
TIME T
TSAMP TS
TS=T+DT

"------CHECK FOR STEADY STATE (SS) AND DEFINE FLAGS
E1 = DELAY(DWr,TW)
E2 = DELAY(E1,TW)
SUM = ABS(DWr)+ABS(E1)+ABS(E2)
NFSS = IF FSS THEN TOEX  ELSE  TOST        "FLAG FOR SS CONDITION
TOEX = IF SUM<TOLEX THEN 1 ELSE 0          " -- TEST OF EXIT
TOST = IF SUM<TOLST THEN 1 ELSE 0          " -- TEST OF START
FSSA = FSS          "AVOID SIMNON PROBLEM WHEN EXPORTING FSS
NCR = IF FSS<1 THEN 0 ELSE IF RT THEN CR+1 ELSE CR
NDC = IF DC < 0.99 THEN COUNT ELSE DC-1
RT  = IF DC < 0.99 THEN 1 ELSE 0           "FLAG  FOR 1ST ITERATION.
```

```
"------DEFINE BASE VALUE FOR Dids (GID), IMPOSING Min. AND Max.
GIDA = C1*Wr/Wrated + C2*TLest/Trated + C3      "PRELIMINARY VALUE.
GID =  IF ABS(GIDA) < TOL4 THEN SIGN(GIDA)*TOL4 ELSE GIDC
GIDC = IF ABS(GIDA) > TOL5 THEN SIGN(GIDA)*TOL5 ELSE GIDA

"------COMPUTE POWER GAIN (1/Pbase) AND SCALED DELTA Pi (DPi)
GPbA = -0.05*Wr/Wrated + 0.0625
GPb = MIN( GPbmax, MAX(GPBmin,GPbA))

NPo  = IF RT THEN Pi ELSE Po                    "Pdc(k-1)
NDPi = IF RT THEN (Pi-Po)*GPb ELSE DPi          "Delta Pdc.

"------EVALUATION OF Cid (CHANGE IN Dids)
"      (A FUZZY RELATION IS USED TO EVALUATE Cid)
" ------PRELIMINARY COMPUTATION OF MEMBERSHIP VALUES
A1 = MIN(1,(-NDPi-P2)/(P1-P2))
A2 = (-NDPi-P3)/(P2-P3)
A3 = (-NDPi/P3)
A5 = NDPi/P3
A6 = (NDPi-P3)/(P2-P3)
A7 = MIN(1,(NDPi-P2)/(P1-P2))

"-------GET INTERVAL INDEX FOR DPi
J  = IF NDPi< -P1 THEN 1 ELSE IF NDPi<-P2 THEN 2 ELSE J1
J1 = IF NDPi<-P3 THEN 3 ELSE IF NDPi< 0  THEN 4 ELSE J2
J2 = IF NDPi< P3 THEN 5 ELSE IF NDPI< P2 THEN 6 ELSE J3
J3 = IF NDPi < P1 THEN 7 ELSE 8

"-------EVALUATION OF DEGREES OF MEMBERSHIP FOR DPi
MP1 = IF J<3 THEN A1 ELSE IF J<4 THEN A2 ELSE IF J<5 THEN A3 ELSE T1
T1 = IF J<6 THEN (1-A5) ELSE IF J<7 THEN (1-A6) ELSE (I-A7)
MP2 = 1-MP1

"-------COMPUTES DEGREE OF MEMBERSHIP FOR LAST Dids
MI1 = IF Dids<-L3 THEN 0 ELSE MIN(1,((Dids+L3)/(L2+L3)))        "POSITIVE
MI2 = IF Dids> L3 THEN 0 ELSE MIN(1,((-Dids+L3)/(L2+L3)))       "NEGATIVE

"-------PERFORM RULE BASE EVALUATION
MRA = MIN(MP1,MI1)
MRB = MIN(MP1,MI2)
MRC = MIN(MP2,MI1)
MRD = MIN(MP2,MI2)

"-----GET CONTROL SIGNAL FOR EACH RELEVANT FUZZY RULE
DIA = IF J<3 THEN I1 ELSE IF J<4 THEN I2 ELSE IF J<5 THEN I3 ELSE T2
T2  = IF J<6 THEN 0.0 ELSE -I3
DIB = -DIA                                      "FOR THIS SPECIFIC RB
DIC = IF J<3 THEN I2 ELSE IF J<4 THEN I3 ELSE IF J<5 THEN 0.0 ELSE T3
T3  = IF J<7 THEN -I3 ELSE -I2
DID = -DIC                                      "FOR THIS SPECIFIC RB

"--------EVALUATE Cid USING HEIGHT DEFUZZIFICATION METHOD
FV = (MRA*DIA+MRB*DIB+MRC*DIC+MRD*DID)/(MRA+MRB+MRC+MRD)
Cid = IF NCR<1 THEN 0 ELSE IF NCR<2 THEN -1.0 ELSE FV

"-------CALCULATION OF NEW Dids
NDids = IF RT THEN Cid ELSE Dids
DAUX = NDIDS*GID
DIDEC = IF ABS(DAUX)<TOL2 THEN SIGN(DAUX)*TOL2*FSS*RT ELSE DAUX

"------PARAMETERS
"DT:5E-06                  "SAMPLING INTERVAL OF THE DISCRETE SYSTEMS
```

```
TW:1E-03                    "PERIOD FOR TEST OF SS CONDITION

TOLST:1                     "TOL. TO START  STEADY STATE COND. (IN RAD/S)
TOLEX:3                     "TOL. TO EXIT STEADY STATE COND.

C1:1.083                    "COEFFICIENTS FOR GID COMUTATION
C2:-3.047
C3:1.496

GPbmax:0.05                 "MAX AND MIN VALUES FOR POWER GAIN
GPbmin:0.0125

Trated:20.83                "RATED MOTOR TORQUE
Wrated:188.5                "RATED Wr IN MECH. RAD./S
TOL2:0.1                    "MIN. CHANGE FOR Dids (A)
TOL4:0.5                    "MIN. VALUE FOR GID (AMPS)
TOL5:1.5                    "MAX. VALUE FOR GID (AMPS)
COUNT=0.5/DT                "DOWN COUNTER INITIAL VALUE (FOR Teff=0.5 S)

"------PARAMETERS FOR FUZZY SETS
P1:1                        "CONSTANTS USED IN DEFINITION OF FUZZY SETS FOR DPi
P2:0.5
P3:0.3
I1:1.0                      "CONSTANTS USED IN DEFINITION OF FUZZY SETS FOR DIds
I2:0.7
I3:0.40
L1:1.0                      "CONSTANT USED IN DEFINITION OF FUZZY SETS FOR LDids
L2:0.1
L3:0.001

"----- INITIAL VALUES
DC:100000                   "DOWN COUNTER
END



" ------------------------------------------------------------------------------------------
"           FEED FORWARD TORQUE COMPENSATION  ROUTINE
" ------------------------------------------------------------------------------------------
DISCRETE SYSTEM FFTC

INPUT iqecp Dids RT FSS AT
OUTPUT Wsl ideca iqec TLest
STATE idec  fdre  iqst  iqsc  fdr
NEW  Nidec Nfdre Niqst Niqsc Nfdr
TIME T
TSAMP TS
TS=T+AT

"-------DEFINES idec, Lm AND STEADY-STATE ESTIMATED D-AXIS FLUX
aidec =  idec+Dids
Lm = Lm0-Mbs2*(aidec-idec0)
Nfdre = if RT then lm*aidec else fdre

"-------COMPUTE diqs(DELTA iqs) , TEST FOR ISmax
"       AND COMPUTE iqst (STAIR-CASE iqs)
diqsa = FSS*(fdre-Nfdre)*iqec/Nfdre
aiqec = iqecp+iqst+diqsa
IP2 = aidec*aidec+aiqec*aiqec                   "SQUARE OF NEW STATOR CURRENT VECTOR
diqs = if IP2 < IMAX then diqsa else 0
"      (IMAX IS THE SQUARE OF MAX. STATOR. CURR. VECTOR)
Niqst = if RT then iqst+diqs else iqst          "Accumulated diqs.
```

```
"-------COMPUTE iqsc (compensatig signal) and iqec (iqs*)
Niqsc = iqsc + AT*(iqst-iqsc)/Talr                    "Filter of rotor time constant.
Nidec = if FSS<1 then idecr else if IP2<IMAX and RT then aidec else idec
ideca = Nidec                                         "Ids*(k).
iqec = min((iqecp+iqsc),sqrt(IMAX-idec*idec))         "Iqs*(k).

"-------EVALUATE SLIP FREQUENCY Wsl, USING ESTIMATED FLUX
Talr=Lr/Rr
fdrc = if FSS then fdre else fdrr                     "Estimate of the new steady state
Nfdr = fdr + AT*(fdrc-fdr)/Talr                       "    value of fdr
Wsl=(Lm*Rr*iqec)/(Lr*fdr)

"--------ESTIMATE MACHINE TORQUE
TLest = Kt*iqec*fdr

"--------PARAMETERS
IMAX = if FSS then IMAX1 else IMAX2
Lr=Lm+Llr                                  "Rotor inductance.
Llr:0.00213                                "Rotor leakage inductance.
Rr:0.478                                   "Rotor resistance.
Lm0:0.06277                                "Magnetizing inductance (unsaturated).
Mbs2:2.588e-03                             "Saturation coefficient.
idec0:4.8  "
idecr:9.786                                "RATED idec
fdrr:0.4833                                "RATED flux
IMAX1:360                                  "MAX. STEADY STATE STATOR CURRENT (SQUARED)
IMAX2:808                                  "MAX. TRANSIENT STATOR CURRENT (SQUARED)
Kt:2.655   "Torque constant.

"-------INITIAL CONDITIONS
fdr:0.4833                                 "INITIAL CONDITION FOR RATED Ids*.
idec:9.786
fdre:0.4833
END


" ------------------------------------------------------------------------------------------
"    REFERENCE CURRENT GENERATOR AND VECTOR ROTATOR
" ------------------------------------------------------------------------------------------
"
CONTINUOUS SYSTEM VECROT
INPUT iqec idec Wr Wsl
OUTPUT iac ibc icc We coswt sinwt
STATE TH
DER DTH
"
TIME T
"--------GENERATE UNIT VECTORS
We=(pole/2)*Wr+Wsl
Dth=We
the=MOD(th,p2)
coswt=cos(the)
sinwt=sin(the)

"--------SYNCHRONOUS TO STATIONARY D-Q REF. FRAME
iqsc=iqec*coswt+idec*sinwt
idsc=idec*coswt-iqec*sinwt

"--------STATIONARY TO ABC REF FRAME
iac=iqsc
ibc=-(idsc*sqrt(3)+iqsc)*0.5
icc=-(iac+ibc)
```

```
"--------DEFINE CONSTANTS
p2=8*ATAN(1)
pole:4
END


" ------------------------------------------------------------------------------------------
"            COMPUTATION OF AVERAGE LOSSES, INPUT AND
"                    OUTPUT POWERS AND EFFICIENCY
" _____------------------------------------------------------------------------------------
CONTINUOUS SYSTEM ECALC
INPUT  Plinv Pild Pin  Po  Pior Piir
OUTPUT Piav
STATE  X1    X2   X3    X4    X5    X6
DER    DX1 DX2 DX3 DX4 DX5 DX6
TIME T

"-------COMPUTE AVERAGE (FILTERED) VALUES OF LOSSES AND POWERS
DX1 = X2
DX2 = A21*X1 + A22*X2 + Piir
Pair = -A21*X1                          "AVE. INPUT POWER OF DIODE BRIDGE
"
DX3 = X4                                "AVE. OUTPUT POWER OF DIODE BRIDGE
DX4 = A21*X3 + A22*X4 + Pior
Paor = -A21*X3
"
DX5 = X6                                "AVE. MACHINE INPUT POWER.
DX6 = A21*X5 + A22*X6 + Pin
Paim = -A21*X5

Piav = Paor
Nsys = 100*Po/Pair
Ndc = 100*Po/Paor                       "Efficiency of IM+INVERTER
Nim = 100*Po/Paim

"-------PARAMETERS AND INITIAL VALUES.
Tal2:40e-03
A21 = -1/(Tal2*Tal2)
A22 = -2/Tal2

X1:2.0368
X3:2.0368
X5:2.0368
END


" ------------------------------------------------------------------------------------------
"              CONNECTING SYSTEM FOR VECTOR CONTROL
" ------------------------------------------------------------------------------------------
CONNECTING SYSTEM CIVCS
"
VAS[RCFL]=VAS[AC]
VBS[RCFL]=VBS[AC]
VCS[RCFL]=VCS[AC]
VA[IMS3]=VA[INVL]
VB[IMS3]=VB[INVL]
VC[IMS3]=VC[INVL]
Vd[INVL]=Vd[RCFL]
"
IA1[INVL]=IA[IMS3]
IB1[INVL]=IB[IMS3]
IC1[INVL]=IC[IMS3]
```

```
IAC[INVL]=IAC[VECROT]
IBC[INVL]=IBC[VECROT]
ICC[INVL]=ICC[VECROT]
IQECP[FFTC]=IQECP[WCTRL]
IQEC[VECROT]=IQEC[FFTC]
IDEC[VECROT]=IDECA[FFTC]
I2[RCFL]=I2[INVL]
DIDS[FFTC] = DIDEC[FEOPT]
"
We[IMS3] = We[VECROT]
Wr[WCTRL] = Wr[IMS3]
Wsl[VECROT] = Wsl[FFTC]
Wr[VECROT] = Wr[IMS3]
We[INVL] = We[VECROT]
Wr[FEOPT] = Wr[IMS3]
DWr[FEOPT] = E[WCTRL]
"
Nsaux[IMS3] = Nsaux[INVL]
RT[FFTC] = RT[FEOPT]
FSS[FFTC] = FSSA[FEOPT]
TLest[FEOPT] = TLest[FFTC]
"
Pi[FEOPT] = Piav[ECALC]
Pior[ECALC] = Pior[RCFL]
Piir[ECALC] = Piir[RCFL]
Pin[ECALC] = Pin[IMS3]
Po[ECALC] = Po[IMS3]
Plinv[ECALC] = Plinv[INVL]
Pild[ECALC] = Pild[RCFL]
"
SINWT[IMS3]=SINWT[VECROT]
COSWT[IMS3]=COSWT[VECROT]

AT[INVL] = 1E-06                    "SPECIFY THE SAMPLING PERIOD
AT[FFTC] = 1E-06
DT[FEOPT] = 1E-06
END



" ----------------------------------------------------------------------------
"                MACRO FOR VECTOR CONTROL SYSTEM
" ----------------------------------------------------------------------------
"
MACRO MIVCS
SYST AC RCFL INVL IMS3 WCTRL VECROT FEOPT FFTC ECALC CIVCS
STORE  iqec[vecrot] idec[vecrot] spd Po[IMS3] Nsys Ndc[ECALC] Nim
STORE  fdr[ims3] fqr[ims3]  Paor  -ADD

split 2 1
plot idec[vecrot]
axes v 0  10 h -0.3 6
text 'ids* for efficiency optimization'
"
" ----- PARAMETER DEFINITION AND INITIAL VALUES
par wref:900
init w:188.5                        "for IMS3
init x:1.6                          "for wctrl
init x1:0.3                         "for ecalc
init x3:0.3                         " ,,
init x5:0.3
SIMU -0.30 4 0.5E-06 / fig9a 0.005
```

```
split  2 1
area 1 1
axes v 0 10 h 0 4
show idec
area 2 1
axes v 0 2000
show Paor
END
```

A.3 Slip Gain Tuning Programs

```
"-------------------------------------------------------------------------------
"              IDEAL HYSTERESIS BAND PWM INVERTER
"-------------------------------------------------------------------------------
DISCRETE SYSTEM INV
INPUT ia ib ic iac ibc icc
OUTPUT va vb vc
STATE A B C
NEW NA NB NC
TIME T
TSAMP TS
TS=T+AT

"------GET CURRENT ERRORS
iae=iac-ia
ibe=ibc-ib
ice=icc-ic

"------CALCULATE NEW STATES OF THE SWITCHES
NA=IF iae>L THEN 1 ELSE IF iae<-L THEN 0 ELSE A
NB=IF ibe>L THEN 1 ELSE IF ibe<-L THEN 0 ELSE B
NC=IF ice>L THEN 1 ELSE IF ice<-L THEN 0 ELSE C
Vd3=Vd/3
va=(2*NA-NB-NC)*Vd3
vb=(2*NB-NA-NC)*Vd3
vc=(2*NC-NB-NA)*Vd3

" ---------- PARAMETERS AND INITIAL STATES.
AT:2E-06                    "Sampling interval.
L:1.2                       "Hysteresys band.
Vd:310                      "Assumed dc link voltage
A:1
B:1
C:0
END


"-------------------------------------------------------------------------------
"      SYNCHRONOUS REFERENCE FRAME D-Q MODEL OF
"      AN INDUCTION MACHINE IN TERMS OF CURRENT
"-------------------------------------------------------------------------------
CONTINUOUS SYSTEM machine
INPUT  va vb vc we coswt sinwt
OUTPUT ia ib ic wr Vds Vqs
STATE iqs  ids  iqr  idr  w  x
DER   diqs dids diqr didr dw dx
TIME t
```

```
pi = atan(1)*4
p2 = pi*2
sqr3 = sqrt(3)
spd = w/pole/pi*60

"----- STATE EQUATIONS
diqs = (lr*(vqs-rs1*iqs)-fse*ids+lm*rr*iqr-lrm*w*idr)/k1
dids = (fse*iqs+lr*(vds-rs1*ids)+lrm*w*iqr+lm*rr*idr)/k1
diqr = (lm*(rs1*iqs-vqs)+w*lsm*ids-iqr*rr*ls+fre*idr)/k1
didr = (-w*lsm*iqs+lm*(ids*rs1-vds)-fre*iqr-ls*rr*idr)/k1
dw = (te-tl-b*w*2/pole)/j*pole/2

tl=kll*wr*wr
wr=w*2/pole                          " SPEED IN MECH. RAD./S

"----- VOLTAGE TRANSFORMATIONS
vqss = va                            "Stationary frame.
vdss = (vc-vb)/sqr3
vqs = vqss*coswt-vdss*sinwt          "Synchronous frame.
vds = vqss*sinwt+vdss*coswt

delw = we-w                          "Slip frequency.
fse = lsr*we-delw*lmm                "Auxiliary expressions.
fre = we*lmm-delw*lsr
te = 0.75*pole*lm*(idr*iqs-iqr*ids)  "Developed torque.

"----- NEW VARIABLES BEING INTRODUCED
DX = (Te - X)/TAL
Tefpu = X/Tb                         "Per-unit filtered torque.

fdr = Lr*idr+Lm*ids                  "D and Q axis rotor flux.
fqr = Lr*iqr+Lm*iqs
frpu = sqrt(fdr*fdr+fqr*fqr)/frb     "Per-unit rotor flux.

" ------ GET MACHINE PARAMETERS
ls = (xm+xs)/wb                      "Stator and rotor inductances.
lr = (xm+xr)/wb
lm = xm/wb                           "Magnetizing inductance.
lrm = lr*lm                          "Auxiliary parameters.
lsm = ls*lm
lsr = ls*lr
lmm = lm*lm
k1 = lsr-lmm

"----- CURRENT TRANSFORMATIONS
iqss = iqs*coswt+ids*sinwt           "Stationary D-Q frame.
idss = ids*coswt-iqs*sinwt
ia = iqss                            "ABC frame.
ib = -(0.5*iqss)-(0.5*sqr3*idss)
ic = -(0.5*iqss)+(0.5*sqr3*idss)

"----- PARAMETERS: FOR 5 HP IM OF AC 200 SYSTEM
wb:377                               "Rated frequency (rad./sec.)
rs1:0.177                            "Stator and rotor resistances.
rr:0.099
xs:0.8068                            "Stator and rotor reactances.
xr:0.5127
xm:12.8105                           "Magnetizing inductance.
j:6.58E-03                           "Machine inertia
b:6.58E-04                           "Viscous damping coefficient
pole:4
```

```
kll=kfactor*3.2027e-04                  "At rated speed and torque, kll=3.2027e-04
kfactor:1                               "Rated conditions
lds:9.216                               "Initial value at rated flux.

TAL:50e-06                              "Filter time constant.
frb:0.3132                              "Base flux
Tb:17                                   "Base torque

END


" ------------------------------------------------------------------------
"                    FUZZY SLIP GAIN TUNER
" ------------------------------------------------------------------------
" NOTE:  USE  E AND CE AS INPUT VARIABLES
"
DISCRETE SYSTEM FSGT
INPUT  DQ DVd Wr Iqec
OUTPUT Kslip                            "Kslip = slip gain

STATE Ks  Err  X1  X2
NEW  NKs  NErr NX1 NX2
TIME T
TSAMP TS
TS = T + AT

" ---------- FUZZY CONTROL 1 - Kf COMPUTATION
MHWr = Min(1,ABS(Wr)/Wb)                "D.M. for HIGH Wr
MLWr = 1 - MHWr                         "       LOW

Iqecpu = Iqec/Iqrated
MHIq = Min(1,ABS(Iqecpu))               "D.M. for HIGH Iqe
MLIq = 1 - MHIq                         "       LOW

" ---------- PERFORM RULE BASE EVALUATION - FUZZY CONTROLLER I
MRA1 = Min(MHWr,MHIq)
MRB1 = Min(MLWr,MHIq)
MRC1 = Min(MHWr,MLIq)
MRD1 = Min(MLWr,MLIq)

" ---------- COMPUTE VALUE OF Kf, USING HEIGHT DEFUZZ. METHOD
PROD = MRA1*Kf2 + MRB1*Kf3 + MRC1*Kf1 + MRD1*Kf2
SUM = MRA1 + MRB1 + MRC1 + MRD1
Kf = PROD / SUM

" ---------- COMPUTE Error AS A COMBINATION OF DQ AND DVd
NErr  = -1*(Kf*DQ + (1-Kf)*DVd)
AUX = NErr*GE                           "Normalized error
E = IF AUX<e1 THEN e1 ELSE IF AUX>e6 THEN e6 ELSE AUX  "Impose bounds

" ---------- COMPUTE CHANGE IN ERROR (ABSOLUTE VALUE)
DE = (NErr-Err)*GC                      "Delta E
NX1 = X1 + (DE - X1)*AT/TAL             "Filtered DE
NX2 = X2 + (X1 - X2)*AT/TAL             "2nd order filter
CE = MIN(c3,ABS(X2))                    "Impose c3 as upper bound for CE

" ---------- DEFINE INTERVAL INDEX (I) AND D.M. FOR CE
I = IF CE<c1 THEN 3 ELSE IF CE<c2 THEN 2 ELSE 1
ci = FUNC(4,I)                          "D.M is obtained by using a function of I
ci1 = FUNC(5,I)
MC1 = (ci - CE) / (ci - ci1)            "Left M.F. for error
MC2 = 1 - MC1                           "Right M.F. for error
```

```
"---------- GET INTERVAL INDEX FOR E
J  = IF E< e2 THEN 1 ELSE IF E< e3 THEN 2 ELSE J1
J1 = IF E< 0 THEN 3 ELSE IF E< e4  THEN 4 ELSE J2
J2 = IF E< e5 THEN 5 ELSE 6

" ---------- DEFINE M.F. FOR E
ei = FUNC(2,J)                          "M.F. definition is given by functions of J
ei1 = FUNC(3,J)

ME1 = (ei - E) / (ei - ei1)             "Left M.F. for error
ME2 = 1 - ME1                           "Right M.F. for error

"-------PERFORM RULE BASE EVALUATION - FUZZY CONTROL 2
MRA = MIN(MC2,ME1)
MRB = MIN(MC2,ME2)
MRC = MIN(MC1,ME1)
MRD = MIN(MC1,ME2)

" ---------- GET Delta Ks VALUES FROM FUNCTION
IJ = (I-1)*7 + J                        "FUNC position of the rule A consequent

DKA = FUNC(1,IJ)                        "Retrieve values of delta Ks, for
DKB = FUNC(1,IJ+1)                      "  the relevant rules.
DKC = FUNC(1,IJ+7)
DKD = FUNC(1,IJ+8)

" ---------- EVALUATE DKs USING HEIGHT DEFUZZIFICATION METHOD
SUM = MRA+MRB+MRC+MRD
DKs = (MRA*DKA+MRB*DKB+MRC*DKC+MRD*DKD)/SUM

" ---------- GET NEW VALUE FOR DELTA Ks
NKs = IF T<0 THEN Ks ELSE Ks+DKs*GKs    "New slip gain
Kslip = Ks

" ---------- PARAMETERS FOR 5 HP MACHINE
Ls: 36.12E-03
Lsig: 3.45E-03
Rs: 0.177
Wb:230.383                              "Wb==2200 RPM
Iqrated:19.11

"---------- FUZZY CONTROLLER  PARAMETERS
e1:-0.6                                 "Define M.F. for Error (E)
e2:-0.3
e3:-0.06
e4:0.1
e5:0.5
e6:1.0
c1:0.2
c2:0.6
c3:1

Kf1:0.8                                 "Define M.F. for Kf
Kf2:0.9
Kf3:0.98

GE:0.5                                  "Input gain for Err
GC:200                                  "Input gain for change in error
GKs:2.5e-03                             "Output gain for DKs

" ---------- INITIAL VALUES
Ks:0.608
```

```
AT:1E-03                        "Sampling interval
TAL:50E-03                      "For CE filter
END



" ---------------------------------------------------------------------------------------------
" COMPUTATION OF FUNDAMENTAL DELTA Q, DELTA Vd
" ---------------------------------------------------------------------------------------------
CONTINUOUS SYSTEM ECOMP
INPUT  We Vds Vqs Idec Iqec
OUTPUT DQE DVdE
STATE  XI   X2   X3
DER    DX1 DX2 DX3
TIME T

"-------COMPUTE FILTERED VALUES
DX1 = (Vds - X1)/TAL            "X1 = Vds filtered
DX2 = (Vqs - X2)/TAL            "X2 = Vqs filtered
DX3 = (X1 - X3)/TAL             "X3 = 2nd order filtered Vds

QE = X2*Idec-X3*Iqec            "Estimated Q, from Idec, Iqec
QEC = We*(Ls*Idec*Idec+Lsig*Iqec*Iqec)"Reference from command signals
Qb = 0.5 + QEC                  "Base value

" ----------- COMPUTE DELTA REACTIVE POWER
DQE = (QEC - QE)/Qb             "Delta Q

" ----------- COMPUTE DELTA Vd
VdEC = rs*Idec - We*Lsig*Iqec   "Command Vds
Vb = 0.1 + ABS(We)*Lsig*ABS(Iqec)  "Base Vds
DVdE = -(VdEC - X3)/Vb          "Delta Vds.

"-------PARAMETERS
Tal:3e-03                       "Filter time constant.
Ls: 34.877E-03                  "Stator inductance.
Lsig: 3.2975E-03                "Modified leakage inductance
Rs: 0.177                       "Stator resistance.
END

" ---------------------------------------------------------------------------------------------
"          CONNECTING SYSTEM FOR SLIP GAIN TUNING
" ---------------------------------------------------------------------------------------------
CONNECTING SYSTEM CMACH

VA[MACHINE]=VA[INV]
VB[MACHINE]=VB[INV]
VC[MACHINE]=VC[INV]
Vds[ecalc]=Vds[MACHINE]
Vqs[ecalc]=Vqs[MACHINE]

IA[INV]=IA[MACHINE]
IB[INV]=IB[MACHINE]
IC[INV]=IC[MACHINE]
IAC[INV]=IAC[VECROT]
IBC[INV]=IBC[VECROT]
ICC[INV]=ICC[VECROT]
IQEC[VECROT]=IQECP[WCTRL]
Iqec[FSGT]=Iqecp[WCTRL]
IDEC[VECROT]=IDEC[WCTRL]
IQEC[ECALC]=IQECP[WCTRL]
```

```
IDEC[ECALC]=IDEC[WCTRL]
"
We[MACHINE]=We[VECROT]
We[ECALC]=We[VECROT]
Wr[WCTRL]=Wr[MACHINE]
Wr[VECROT]=Wr[MACHINE]
Wr[FSGT] = Wr[MACHINE]
Wsl[VECROT]=Wsl[WCTRL]
"
SINWT[MACHINE]=SINWT[VECROT]
COSWT[MACHINE]=COSWT[VECROT]

DQ[FSGT] = DQE[ECALC]
DVd[FSGT] = DVdE[ECALC]
Ks[WCTRL] = Kslip[FSGT]
"
END


"-------------------------------------------------------------------------
"                          MACRO FOR  SLIP GAIN TUNING
"-------------------------------------------------------------------------
MACRO MMACH
SYST INV MACHINE WCTRL VECROT  ECOMP FSGT CMACH
STORE iqecpu Tefpu frpu spd
PLOT fct

error 0.0001
IMPORT   MYFUNC < MYFUNC /2
par AT[INV]:5E-06
split 3 2

INIT Ks:0.304
par wref:1 100

par kfactor:4.0
par GC:500
par GKs:0.3e-03

AREA 1 1
axes h -0.5 2 v 0 2
TEXT 'Ks/Kso'
SIMU -0.50 2  /fig11B  0.001 //MYFUNC

area  1 1
ashow  0 2  iqecpu
TEXT 'iqs*(pu)'
area 2 1
ashow  0  2 tefpu
text  'Te (pu)'
area 3 1
ashow 0 2 frpu
text 'Fr (pu)'

END
```

APPENDIX B

TMS320C25 ASSEMBLY PROGRAMS LISTING

B.1 Efficiency Optimization and Vector Control Programs

```
;-------------------------------------------------------------------------
;   INDIRECT VECTOR CONTROL WITH FUZZY LOGIC
;         EFFICIENCY OPTIMIZATION PROGRAM
;-------------------------------------------------------------------------
; GILBERTO C. D. SOUSA
; UNIVERSITY OF TENNESSEE, KNOXVILLE.
;
; THIS  PROGRAM INCLUDES:
;          1 - VECTOR CONTROL FUNCTIONS
;          2 - SPEED COMPUTATION ROUTINE
;          3 - PI SPEED CONTROL
;          4 - ADC ISR FOR TWO CHANNELS
;          5 - DAC ISR
;          6 - DC LINK POWER CALCULATION
;          7 - SLIP GAIN COMPUTATION FOR VARIABLE FLUX
;          8 - FEED-FORWARD TORQUE COMPENSATION
;          9 - FUZZY EFFICIENCY OPTIMIZATION CONTROLLER
;          10 - TRANSITION CONTROL
;
;NOTE: USING DATA FROM 5 hp CLASS B MACHINE.
;
;-------------------------------------------------
;              I/O PORT ADDRESS
;-------------------------------------------------
MCNT:          EQU    0       ;Encoder pulse count.
TCNT:          EQU    1       ;Clock pulse count.
INSTU:         EQU    2       ;Inverter status READ.
INCOM:         EQU    2       ;Inverter command WRITE.
DATA:          EQU    8       ;Buffer data port.
ADDR:          EQU    9       ;Buffer address port.
READHOST:      EQU    10      ;Host read port.
REFRESH:       EQU    11      ;Enable buffer refresh.
DAC:           EQU    13      ;DAC latch.
HOSTSTAT:      EQU    14      ;Host status.
WRITEHOST:     EQU    15      ;Host write port.

;-------------------------------------------------
;                  CONSTANTS
;-------------------------------------------------
PRDC:          EQU    2000    ;Main sampling time (=0.2 msec).
Iderated:      EQU    12826   ;Rated Ids = 9.786 A (= 12826).
SLIPGAIN:      EQU    810     ;Rated slip gain.
NUMKSLC:       EQU    179     ;Numerator of SG (=LmRr/Lr).1 PU = 179
DT:            EQU    4000    ;Delta t for theta calculation.
TBL90:         EQU    512     ;90 degree offset.
SSAMPC:        EQU    5       ;Speed sampling time counter. (=1msec.)
```

```
STMAXC:         EQU     25      ;Max. speed sampling time. (5 ms)
DELWRCC:        EQU     32767-100 ; Max speed variation in 1 ms.

; ---------- MEMORY MAPPED REGISTERS (ALREADY DEFINED IN RESMON)
;DRR:           EQU     0       ;A/D input data register
;TIM:           EQU     2       ;Timer.
;PRD:           EQU     3       ;Period register.
;IMR:           EQU     4       ;Interrupt mask register.
;GREG:          EQU     5       ;Global memory allocation register.
;
; ---------- DEFINE CONSTANTS FOR PI CONTROLLER
K1C:            EQU     3929    ;3935   ;For kp=0.4 and ki=0.8: K1C=3935
K2C:            EQU     -3920   ;               : K2C=-3927
IQMV:           EQU     322     ;|Iqe| limit / 64 == 322
                                ;Maximum value of Iqs*=16.228 A (= 21270)

; ---------- DEFINE CONSTANTS FOR FIRST ORDER FILTER1 (FOF)
;         FOR SLIP GAIN COMPUTATION AND FFTC.
;         CURRENTLY USING Tr = 0.10778 s.
A1C:            EQU     32465
B1C:            EQU     504
C1C:            EQU     19570
D1C:            EQU     151

; ---------- DEFINE CONSTANTS FOR FOF OF Iqs*, for FFTC only!
A2C:            EQU     32442
B2C:            EQU     503
C2C:            EQU     21127
D2C:            EQU     163

; ---------- DEFINE CONSTANTS FOR FOF OF Pdc.
A3C:            EQU     32702
B3C:            EQU     307
C3C:            EQU     6983
D3C:            EQU     32

; ---------- CONSTANTS FOR FFTC ROUTINE.
Idec0:          EQU     6302    ;Linear region breakpoint.
Mbs2c:          EQU     21200   ;Coefficient for Lm computation (1.216e-03).
Lm0c:           EQU     20568   ;Linear value of Lm (0.06277 H).
Idmaxc:         EQU     17038   ;(32767-Idmax), Idmax=15729 (12A).
Idminc:         EQU     18350   ;(32768+Idmin)/2, Idmin=3932 (3A).
IMAX:           EQU     18830   ;Square of max. stator cur.(Imax=18.95^2).

; ---------- CONSTANTS FOR TRANSITION CONTROL.
TOLex:          EQU     300     ;Limit to exit eff. opt. mode.
TOLst:          EQU     100     ;Limit to start eff. opt. mode.

; ---------- DEFINE CONSTANTS FOR FEOPT ROUTINE.
I1:             EQU     32767   ;Definition of fuzzy sets for DIds (Delta
I2:             EQU     22938   ;  Ids* ).
I3:             EQU     13107
L3:             EQU     328     ;Def. of fuzzy sets for LDIds (Last DIds).
P1:             EQU     2154    ;Def. of fuzzy sets for DPi (Delta Pi).
P2:             EQU     1077
P3:             EQU     646

DPimaxc:        EQU  32767-100  ;Used to clamp (Pi(k)-Pi(k-1) to 100
                                                ;  (43W).
L2limc:         EQU  32767-3277 ;Clamp LDids to 3277.

CG1C:           EQU     2955    ;C1, C2 and C3 coef. for GID comp.
```

```
CG2C:           EQU     -31827
CG3C:           EQU     1961
G1C:            EQU     -3438     ;Coefficients for GPb computation.
G2C:            EQU      5408
GIDMAXC:        EQU 32767-1966    ;Const. for max and min GID values.
GIDMINC:        EQU     16580     ; == (32768+393)/2, min GID = 393.
GPBMAXC:        EQU 32767-6760    ; Const. for max and min GPb comp.
GPBMINC:        EQU     18356     ; == (32768+3943)/2, min GPb=3943.
DIdminc:        EQU     131       ;Min. value for DIde (==0.1 A).

RPI2C:          EQU     7606      ;1/(P1-P2).
RP23C:          EQU     19007     ;1/(P2-P3).
RP3C:           EQU     12681     ;1/P3.
RL23C:          EQU     9089      ;1/(L2+L3).

; ------------------------------------------------------
;               DEFINE DATA RAM
; ------------------------------------------------------
; ---------- PAGE 0  SSV0 and SSV1 must be in PAGE 0 !
                ORG     061H
SSV0:           DS      1
SSV1:           DS      1
SSTEMP:         DS      1

; ---------- PAGE 6
                ORG     00H       ;On-board data RAM (Block B1).
OUTBFA:         DS      1         ;Channel A output buffer.
OUTBFB:         DS      1         ;Channel B output buffer.
COMMAND:        DS      1         ;Host command word.
COMMAND2:       DS      1         ;Value passed to TMSC25.
DISP:           DS      1         ;Value passed to HOST.
HOSTBUF:        DS      1         ;Buffer host status.
DUMMY:          DS      1         ;Dummy register.

Vd:             DS      1         ;DC link voltage.
Id:             DS      1         ;DC link current.

;---------- MAIN INTERRUPT (0.2 ms) ROUTINE.
THETAL:                 DS      I
THETAH:         DS      1
We:             DS      1
Kslip:          DS      1
SIN:            DS      1
COS:            DS      1
Iqs:            DS      I
Ids:            DS      1
Ias:            DS      1
Ibs:            DS      1
RT34:           DS      1
TEMP0:          DS      1

;---------- DATA BUFFER STORAGE VARIABLES.
BUFAL:          DS      1
BUFAH:          DS      I
COL:            DS      1
LOW:            DS      1
BFADATA:        DS      1
BFBDATA:        DS      1
TEMP2:          DS      1
SACCH:          DS      1
SACCL:          DS      1
ZERO:           DS      1
```

```
ONE:            DS      1

; ---------- FEEDBACK SPEED COMPUTATION ROUTINE.
Wrtst:          DS      1
STP:            DS      1
SSAMP:          DS      1
MNEW:           DS      1
MOLD:           DS      1
TNEW:           DS      1
TOLD:           DS      1
Wr1:            DS      1
WROLD:          DS      1
DELM:           DS      1
DELT:           DS      1
DELWRC:         DS      1

;---------- DEFINE VARIABLES FOR PI CONTROLLER
IQH:            DS      1
IQL:            DS      1
ILIM:           DS      1       ;Limit for test of iqs*max (32767-IQMAX).
EK:             DS      1       ;Speed error e(k).
EK1:            DS      1       ;e(k-1).
WREF:           DS      1       ;Speed reference.
KWREF:          DS      1       ;Constant for RPM-->decimal convertion.
KBR:            DS      1       ;  "     " decimal -->RPM convertion.
IQMAX:          DS      1       ;IQMAX location.
K1:             DS      1       ;Modified PI gains.
K2:             DS      1

; ---------- VARIABLES FOR HOST/DSP COMMUNICATION.
INCOMDATA:      DS      1
MSTEP:          DS      1       ;Main routine step.
VAR1:           DS      1       ;Buffer storage var. 1 (Ch. A).
VAR2:           DS      1       ;Buffer storage var. 2 (Ch. B).

; ---------- VARIABLES FOR TRANSITION CONTROL.
FSS:            DS      1       ;Flag for steady state (1=ss, 0=dyn.mode).
CSS:            DS      1       ;Counter for state transition.
EOM:            DS      1       ;User defined Eff. Opt. Mode start/stop.

; ---------- VARIABLES FOR NEW ADC.
PHIGH:          DS      1
PLOW:           DS      1
TREG:           DS      1
Idc:            DS      1
SPdH:           DS      1       ;Sum of Vd(k)*Idc(k)
SPdL:           DS      1

; ---------- VARIABLES FOR Pdc COMPUTATION
SWT0:           DS      1       ;S/W timer 0: variable Ts, for Buffer storage
SWT1:           DS      1       ;S/W timer 1: 1 ms for PI speed and FFTC filter
SWT2:           DS      1       ;S/W timer 2: 50 ms for Pdc ave comp.
SWT3:           DS      1       ;S/W timer 3: 0.5 s for FEOPT, FFTC
Pdc:            DS      1       ;Pdc average over 50 ms
Pdck:           DS      1       ;Instantaneous Pdc(k)
TEMP:           DS      1       ;Used in ADCGIL
PTEMP:          DS      1       ;Used in computation of Pdck

; ---------- VARIABLES FOR SLIP GAIN COMPUTATION (VARIABLE FLUX)
Idef:           DS      1       ;Filtered value of Ide (=flux/Lm)
xks:            DS      1       ;State variable for filter 1
A1:             DS      1       ;Constants for filter 1
```

```
B1:             DS      1
C1:             DS      1
D1:             DS      1
NUMKSL:         DS      1       ;Numerator for slip gain computation

; ---------- VARIABLES FOR FFTC
DIde:           DS      1       ;Delta Ide
AIde:           DS      1       ;Auxiliary Ide
Ide:            DS      1       ;Ids*
Lm:             DS      1       ;Magnetizing inductance
Mbs2:           DS      1       ;Coefficient for Lm computation
Nfdre:          DS      1       ;New steady state estimate of rotor flux
fdre:           DS      1       ;Current value of flux estimate
DIqe:           DS      1       ;Delta Iqe
Iqep:           DS      1       ;Iqs*', from PI speed controller
AIqe:           DS      1       ;Actual Iqs*, including compensation
Iqe:            DS      1       ;Auxiliary variable (=Iqe)
SDIqe:          DS      1       ;Sum of Delta Iqs. Staircase shape
SDIqf:          DS      1       ;Output of 1st order filter (filtered SDIqe)
Idmax:          DS      1       ;(32767-Idmax)
Idmin:          DS      1       ;(32768-Idmin)/2
IP2:            DS      1       ;Is**2
Lm0:            DS      1       ;Linear value of Lm
xkf:            DS      1       ;State variable for 1st order filter FOFT
TLest:          DS      1       ;Torque estimate

; ---------- VARIABLES FOR FEEDBACK SPEED SIGNAL
A2:             DS      1       ;Coefficients for filter FOFW
B2:             DS      1
C2:             DS      1
D2:             DS      1
XWK:            DS      1       ;State variable
Wr:             DS      1       ;Filtered speed signal
XKL:            DS      1       ;16 lsb of xkw (Speed fb)
XKLS:           DS      1       ;       xks     (Slip gain)
XKLT:           DS      1       ;       xkf     (Torque comp.)

; ---------- TEST VARIABLES.
SWeH:           DS      1       ;Average computation of Wr
SWeL:           DS      1
Iqave:          DS      1
SPdH2:          DS      1
SPdL2:          DS      1

; ---------- VARIABLES FOR Pdc FILTER
Pdc1:           DS      1       ;Original Pdc
Pd16:           DS      1       ;Pd ave. over 8 samples
Poffset:        DS      1
A3:             DS      1       ;Coefficients for filter FOFP
B3:             DS      1
C3:             DS      1
D3:             DS      1
xpk:            DS      1       ;State variable
XPL:            DS      1       ;16 lsb of xpk (Pdc filter)

; ---------- DEFINE VARIABLES AND ADJUSTABLE CONSTANTS FOR FEOPT.
; ---------- PAGE 7.
                ORG     30h     ;Use data page 7 for FEOPT variables.

TLest2:         DS      1       ;Load torque estimate, dp 7.
Wr2:            DS      1       ;Speed, dp7.
DPi:            DS      1       ;DPi = Pi(k)-Pi(k-1).
```

```
CG1:        DS    1     ;Coef. for GID computation
CG2:        DS    1
CG3:        DS    1
G1:         DS    1     ;Coef. for GPb computation
G2:         DS    1
GID:        DS    1     ;Scale factor for DIds
GIDMAX:     DS    1     ;Impose max. value to GID (32767-1966)
GIDMIN:     DS    1     ;Impose min. value to GID (32767+655)/2

GPb:        DS    1     ;Scale factor for DPi
GPbMAX:     DS    1     ;Impose max. value to GPb (32767-17564)
GPbMIN:     DS    1     ;Impose min. value to GPb (32767+4378)/2
DIdmin:     DS    1     ;DIdmin is the min. value for DIde
JP:         DS    1     ;Interval index for DPi
MP1:        DS    1     ;Degree of membership for DPi (Left)
MP2:        DS    1     ;Same (right)

MI1:        DS    1     ;Degree of membershipt for last DIds (Pos.)
MI2:        DS    1     ;Same (Neg.)
MRA:        DS    1     ;Truth value for rule A
MRB:        DS    1     ;                    B
MRC:        DS    1     ;                    C
MRD:        DS    1     ;                    D

DIA:        DS    1     ;Control signal due to rule A
DIB:        DS    1     ;                      B
DIC:        DS    1     ;                      C
DID:        DS    1     ;                      D
DIds:       DS    1     ;Per-unit value of delta Ids
DMONE:      DS    1     ;Degree of membership value of one

RP12:       DS    1     ; 1/(P1-P2)
RP23:       DS    1     ; 1/(P2-P3)
RP3:        DS    1     ; 1/P3
RL23:       DS    1     ; 1/(L2+L3)
Pik:        DS    1     ; Pdc(k)
Pik1:       DS    1     ; Pdc(k-1)
DPimax:     DS    1
L2lim:      DS    1
SUMMR:      DS    1     ; Sum of truth values of fired rules
SUMPROD:    DS    1     ; Sum of products (sum of MRi*DIi, i=A,B,C,D)
TEMPi:      DS    1     ; Temporary variable

; ------------------------------------------------------------
;         RESET AND INTERRUPT VECTORS
; ------------------------------------------------------------
;         ORG   0
;RESET:   B     INIT          ;Reset vector:currently in RESMON
         ORG   4
INT1:    B     AINT          ;Channel A service interrupt routine
         ORG   6
INT2:    B     BINT          ;Channel B service interrupt routine
         ORG   24
INTT:    B     TINT           ;MAIN INTERRUPT (TIMER)
         ORG   26
RCV:     B     ININT         ;ADC service interrupt routine

; ------------------------------------------------
;             INITIALIZE PROCESSOR
; ------------------------------------------------
         ORG   1024          ;WITH RESMON ORG MUST BE 1024
INIT:    LDPK  0
```

```
        LALK    PRDC
        SACL    PRD             ;SET TIMER PERIOD
        SOVM
        SSXM                    ;Set sign extension mode.
        SPM     0               ;No shift in product register output
        FORT    0               ;Configure for 16 bit word serial data
        SXF                     ;Set output channel flag
        LACK    30              ;Enable RINT(16), AINT(2), BINT(4),TIMER(8)
        SACL    IMR             ;Set interrupt mask register
        LDPK    6
        LACK    1
        SACL    INCOMDATA
        OUT     INCOMDATA,INCOM         ;Clear M and T counters

; ---------- CONSTANT INITIALIZATION.
        ZAC
        SACL    SSAMP
        LALK    DELWRCC
        SACL    DELWRC

; ---------- Initialization of task handler and speed computation
        LDPK    6
        IN      MOLD,MCNT       ;Initialize MOLD=MCOUNT
        IN      TOLD,TCNT       ;   "     TOLD=TCOUNT
        ZAC
        SACL    Wr1
        SACL    Wr
        SACL    Wref
        SACL    TLest
        SACL    WROLD
        SACL    OUTBFA
        SACL    OUTBFB
        SACL    STP
        LALK    SLIPGAIN
        SACL    Kslip
        LALK    28377
        SACL    RT34

; ---------- BUFFER INITIALIZATION
        LACK    3
        SACL    BUFAH
        LALK    0FE00H
        SACL    BUFAL
        ZAC
        SACL    VAR1            ;Selection of variable to be stored
        SACL    VAR2
        SACL    ZERO
        LACK    1
        SACL    ONE

; ---------- FILTER 1 INIT.: SLIP GAIN COMPUTATION UNDER VAR. FLUX
        LALK    A1C             ;Initialize coeficients
        SACL    A1
        LALK    B1C
        SACL    B1
        LALK    C1C
        SACL    C1
        LALK    D1C
        SACL    D1

        LALK    A2C             ;Initialize coeficients for Wr filter
        SACL    A2
```

```
        LALK    B2C
        SACL    B2
        LALK    C2C
        SACL    C2
        LALK    D2C
        SACL    D2

        LALK    A3C             ;Initialize coeficients for Pdc filter
        SACL    A3
        LALK    B3C
        SACL    B3
        LALK    C3C
        SACL    C3
        LALK    D3C
        SACL    D3

        LALK    Iderated        ;Initialize Idef (always to rated value)
        SACL    Idef
        LALK    20133           ;Initialize xks to rated value (due to Ider)
        SACL    xks
        LALK    NUMKSLC         ;Initialize numerator of slip gain fraction
        SACL    NUMKSL

; ---------- INITIALIZE VARIABLES FOR PI CONTROLLER AND SPEED FILTER
        ZAC
        SACL    WR1             ;Original feedback speed
        SACL    WR              ;Filtered feedback speed
        SACL    XWK
        SACL    WREF
        SACL    Iqep
        SACL    Iqe
        SACL    IQH
        SACL    IQL
        SACL    EK1
        LALK    K1C
        SACL    K1
        LALK    K2C
        SACL    K2
        LALK    IQMV
        SACL    IQMAX           ;Store iqmv into IQMAX
        LALK    32767
        SUB     IQMAX   ;
        SACL    ILIM            ;ILIM = 32767 - IQMAX

; ---------- INITIALIZE VARIABLES FOR Pdc COMPUTATION
        LALK    5               ;Multiple of 0.2ms (50 <--> 10 ms)
        SACL    SWT0            ;Counter for Buffer storage (variable)
        LACK  5
        SACL    SWT1            ;Counter for PI speed ctrl and FOF for FFTC
        LALK    1000
        SACL    SWT2            ;Counter for Pdc ave. computation
        LALK    1000            ; 1000bits or 2 sec.
        SACL    SWT3            ;Counter for FEOPT AND FFTC
        ZAC
        SACL    FSS             ;Initially FSS=0 (dynamic mode)
        SACL    CSS             ;Counter for transition control
        SACL    EOM             ;Eff. Opt. Mode (user defined)
        SACL    Pd16            ;8 samples average
        LARK AR5,7              ;Counter for Pd16 average computation

; ---------- INITIALIZE VARIABLES AND CONSTANTS FOR HOST/DSP COMM.
        LALK    29320
```

```
        SACL    KWREF   ;rpm to binary coef.
        LALK    4578
        SACL    KBR             ;Binary to rpm coef.

; ---------- INITIALIZE VARIABLES AND CONSTANTS FOR FFTC
        LALK    Mbs2c
        SACL    Mbs2
        LALK    Idmaxc          ;Current limits: Idmax = 12 A
        SACL    Idmax           ;DIdmax=32767-Idmax
        LALK    Idminc          ;           Idmin = 3 A
        SACL    Idmin           ;DIdmin=(32768-Idmin)/2
        LALK    Iderated        ;Rated Ide = 9.216 A for 5 hp m/c.
        SACL    Ide
        SACL    AIde            ;AIde is also initialized to Iderated
        LALK    6335            ;Rated flux is 0.4833 Wb (=6335)
        SACL    Nfdre
        SACL    fdre            ;Both fdre and Nfdre are init. to rated values.
        LALK    16340           ;Rated Lm (49.38 mH or 16340 bits)
        SACL    Lm
        LALK    Lm0c
        SACL    Lm0

        ZAC                     ;Initialize to zero the remaining variables
        SACL    SDIqe           ;Sum of delta Iqs*
        SACL    SDIqf           ;Filtered version of SDIqe
        SACL    DIde            ;Delta Ide
        SACL    DIqe            ;Delta Iqe
        SACL    xkf             ;State variable for FOFT
        SACL    TLest           ;Torque estimate
        SACL    DISP

        SACL    XKL             ;Lsb of fof, speed, slip gain , torque comp.
        SACL    XKLS
        SACL    XKLT


; ---------- INITIALIZE VARIABLES AND ADJUSTABLE CONST. FOR FEOPT
        LDPK    7
        LALK    CG1C            ;Coef. for GID computation
        SACL    CG1
        LALK    CG2C
        SACL    CG2
        LALK    CG3C
        SACL    CG3

        LALK    GIDMAXC         ;Max and min values for GID
        SACL    GIDMAX
        LALK    GIDMINC
        SACL    GIDMIN

        LALK    G1C             ;Coef. for GPb computation
        SACL    G1
        LALK    G2C
        SACL    G2

        LALK    GPbMAXC         ;Max and min values for GPb
        SACL    GPbMAX
        LALK    GPbMINC
        SACL    GPbMIN

        LALK    DIdminc         ;Min value for DId
        SACL    DIdmin
```

```
        LALK    1000
        SACL    DMONE           ;Value for degree of membership one
        LALK    DPimaxc         ;Value used in the DPi max computation
        SACL    DPimax
        LALK    L2limc          ;Used for max LDIds computation
        SACL    L2lim
        ZAC
        SACL    GID

; ---------- Initialize constants for evaluation of degree of memb.
        LALK    RP12C
        SACL    RP12
        LALK    RP23C
        SACL    RP23
        LALK    RP3C
        SACL    RP3
        LALK    RL23C
        SACL    RL23

; ------------------------------------------------------------------------
; BACKGROUND LOOP - HOST COMMUNICATION AND IDLE TIME
; ------------------------------------------------------------------------
MAIN:   EQU     $
        EINT
        LDPK    6               ;Write selected variable to HOST
        OUT     DISP,WRITEHOST  ;VAR ---> WRITEHOST (PA15)

; ------------------------------------------------------------------------
;           HOST/DSP16 COMMUNICATION ROUTINE
; ------------------------------------------------------------------------
NCLUD HOSTDSP.ASM

; ------------------------------------------------------------------------
;                   MAIN INTERRUPT ROUTINE
; ------------------------------------------------------------------------
;---------- CONTEXT SAVE
;               AR7 IS THE STACK POINTER ( AR7=124!)
TINT:   LARP    AR7
        LARK    AR7,124
        MAR     *-
        SSTI    *-
        SST     *-
        EINT
        SACH    *-
        SACL    *-
        SPH     *-
        SPL     *-
        MPYK    1
        SPL     *-
        LDPK    6

; ------------------------------------------------------------------------
;           PERFORM FEEDBACK SPEED COMPUTATION
; ------------------------------------------------------------------------
        CALL    SPEED
        LAC     Wrl
        SACL    Wr
; ------------------------------------------------------------------------
;           PI SPEED CONTROL
;           FIRST ORDER FILTERS FOR SLIP GAIN AND FFTC
;           TRANSITION CONTROL ROUTINE
; ------------------------------------------------------------------------
```

```
TS1MS:  LDPK    6               ;1 ms sampling time routines
        LAC     SWT1            ;Check for the sampling time
        SUBK    1
        SACL    SWT1            ;Decrement counter SWT1
        BGZ     TS1RET
        LACK    5               ;5<-->1ms
        SACL    SWT1            ;Reload counter if zero
        CALL    PI              ;PI speed controller
        CALL    FOF1            ;Compute estimate of "flux" (Idef).
        CALL    FOFT            ;Get compensated Iqe, for FFTC
        CALL    FOFW            ;Filter for Iqs* (FFTC use only).
        CALL    TRCTL           ;Execute transition control routine
        CALL    FOFP            ;Compute filtered Pdc.
        CALL    AVE16           ;Average variables for monitoring only.
TS1RET: EQU     $

;-------------------------------------------------------------------
;       PERFORM BASIC VECTOR CONTROL FUNCTIONS
;-------------------------------------------------------------------
        CALL    VECROT          ;Original program: slip calc. + v.r

;-------------------------------------------------------------------
;     OUTPUT VARIABLES (FOR DAC)
;-------------------------------------------------------------------
        SPM     1
        LT      Ias             ;Compensates for DAC gain mismatches
        MPYK    203
        PAC
        ADDH    Ias
        SACH    Ias
        LAC     Ias

        ADDK    90              ;Offset compensation for DAC A
        SACL    OUTBFA          ;Channel A of DAC

        LT      Ibs             ;Compensates for DAC gain mismatches
        MPYK    128
        PAC                     ;PAC is the correct command here.
        ADDH    Ibs
        SACH    Ibs
        LAC     Ibs
        ADDK    120             ;Offset compensation for DAC B
        SACL    OUTBFB          ;Channel B of DAC

;-------------------------------------------------------------------
;             COMPUTATION OF Pdc AVERAGE
;-------------------------------------------------------------------
        LDPK    6
        LAC     SWT2            ;Check for the sampling time of Pdc comput.
        SUBK    1
        SACL    SWT2            ;Decrement counter SWT2
        BGZ     PDCRET
        LALK    1000            ;SWT2 = 200 ms (1000 x 0.2 ms)
        SACL    SWT2            ;Reload counter if zero and
        CALL    PDCAVE          ;Compute Pdc ave.
PDCRET: EQU     $

;-------------------------------------------------------------------
;       FUZZY EFFICIENCY OPTIMIZATION AND
;       FEED-FORWARD TORQUE COMPENSATION
;-------------------------------------------------------------------
        LAC     FSS             ;Test operating mode
```

```
        BGZ     SSMD            ;If FSS = 1, go to steady state mode
        ZAC                     ;Dyn. mode. Make SDIqe = 0
        SACL    SDIqe           ; ... and bypass FFTC and FEOPT
        B       FRET
SSMD:
        LAC     SWT3            ;Check for the sampling time of FFTC AND FEOPT
        SUBK    1
        SACL    SWT3            ;Decrement counter SWT3
        BGZ     FRET
        LALK    10000           ;TIMER = 2 S (10000*0.2ms)
        SACL    SWT3            ;Reload counter if zero and
        CALL    FEOPT           ;Calculate DIde by fuzzy logic
        CALL    FFTC            ;Compute value of DIqe
FRET:   EQU     $

; ---------------------------------------------------
;               BUFFER STORAGE
; ---------------------------------------------------
; ---------- SELECT THE VARIABLE TO BE STORED IN CHANNEL A
        SPM     1
        LDPK    6                               ;************************************
        LAC     VAR1                            ;*        VAR1 = 1  STORE:  Ide     *
        SUBK    1               ;Case VAR1 = 1  ;*                  2      Iqe     *
        BNZ     VA2                             ;*                  3      Pdc     *
        LAC     Ide                             ;************************************
        SACL    DISP
        B       VAEND

VA2:    SUBK    1               ;Case VAR1 = 2
        BNZ     VA3
        LAC     Iqe,1
        SACL    DISP
        B       VAEND

VA3:    SUBK    1               ;Case VAR1 = 3
        BNZ     VBSTART         ;Wrong value for VAR1; check VAR2
        LAC     Pd16            ;Pd16 is the 1.6 s average of Pdc
        SACL    DISP

VAEND:  SACL    BFADATA         ;Store selected variable into CH. A

; ---------- SELECT THE VARIABLE TO BE STORED IN CHANNEL B
VBSTART:
        LDPK    6                               ;************************************
        LAC     VAR2                            ;*VAR2 =          1       Pdc     *
        SUBK    1               ;Case VAR2 = 1  ;*                2       SDIqf   *
        BNZ     VB2                             ;*                3       Wr      *
                                                ;************************************
        LAC     Pdc
        SUB     Poffset
        SACL    DISP,5
        LAC     DISP
        B       VBEND

VB2:    SUBK    1               ;Case VAR2 = 2
        BNZ     VB3
        LAC     SDIqf,1         ;Torque compensating signal.
        SACL    DISP
        B       VBEND

VB3:    SUBK    1               ;Case VAR2 = 3
        BNZ     VAREND          ;Value of VAR2 is not valid; ignore it.
```

```
        LAC     Wr1,1           ;Speed.
        SACL    DISP
VBEND:  SACL    BFBDATA

; ---------------------------------------------------
;   BUFFER STORAGE SUBROUTINE
; ---------------------------------------------------
VAREND:         LDPK    6
        LAC     SWT0            ;Check for the sampling time of Buffer Storage.
        SUBK    1
        SACL    SWT0            ;Decrement counter SWT0
        BGZ     BUFRET          ;By-pass if not required
        LALK    100             ;SWT0 = 10 ms (variable) 50 x 0.2ms = 10ms
        SACL    SWT0            ;Reload counter if zero and
        CALL    BUFSTR          ;Perform buffer storage
BUFRET: EQU     $

;---------- CONTEXT RESTORE
        LARP            AR7
        MAR             *+
        MAR             *+
        LT              *-
        MPYK    1
        LT              *+
        MAR             *+
        LPH             *+
        ZALS            *+
        ADDH    *+
        DINT
        LDPK            0
        SST1            SSTEMP
        BIT             SSTEMP,0bh
        BBNZ    PPPP
        LDPK    6
        LST             *+
        EINT
        RET
PPPP:   LDPK    6
        LST     *+
        EINT
        RET

; ---------------------------------------------------
;  LIST OF INCLUDED SUBROUTINES
; ---------------------------------------------------
NCLUD   ADC.ASM                 ;ADC interrupt service subroutine.
NCLUD   BUFSTO.ASM              ;Buffer storage subroutine
NCLUD   DAC.ASM                 ;DAC interrupt service subroutine.
NCLUD    PI.ASM                 ;PI speed controller.
NCLUD    SPEED.ASM              ;Feedback speed computation.
NCLUD    PDCOMP.ASM             ;Average DC link power computation.

NCLUD   FFTC.ASM                ;Feedforward pulsating torque compensation.
NCLUD   FEOPT.ASM               ;Fuzzy efficiency optimization controller.
NCLUD   FOFSG.ASM               ;First order filter for slip gain computation
NCLUD   FOFT.ASM                ;First order filter for Iqe compensation, FFTC
NCLUD   FOFW.ASM                ;First order filter for Iqs* (Use in FFTC only)
NCLUD   FOFP.ASM                ;First order filter for Pdc.
NCLUD   TRNCTRL.ASM             ;Transition control.
NCLUD SIN.ASM                   ;Table for sin/cos theta computation.
NCLUD VR.ASM                    ;Vector rotation.
NCLUD AVE16.ASM                 ;Average Pdc computation over 1.024 s.
```

```
; ---------- END OF PROGRAM ----------

; ------------------------------------------------------------
;         ADC INTERRUPT SERVICE ROUTINE
; ------------------------------------------------------------
;         AR6 IS USED HERE TO STORE DRR CONTENT
;
ININT: EQU      $
       BIOZ     CHB             ;If channel B jump to CHB
;      ---------- CONTEXT SAVE AND INPUT DRR CONTENT
       SST1     SSV1            ;SST1 and SST instr. always load into page 0!
       SST      SSV0
       LDPK     0
       LAR      6,DRR           ;Note that DRR is in DATA PAGE 0
       LDPK     6               ;Remaining of ISR are in DATA PAGE 6
       SACH     SACCH           ;Save ACC
       SACL     SACCL
       SPM      0
       SPH      PHIGH           ;Save P register
       SPL      PLOW
       MPYK     1               ;Save T register (PR=TR)
       SPL      TREG

CHA:   SAR      6,Vd
       B        CHEXT

CHB: SST1       SSV1            ;SST1 and SST instr. always load into page 0!
       SST      SSV0
       LDPK     0
       LAR      6,DRR           ;Note that DRR is in DATA PAGE 0
       LDPK     6               ;Remaining of ISR are in DATA PAGE 6
       SACH     SACCH           ;Save ACC
       SACL     SACCL
       SPM      0
       SPH      PHIGH           ;Save P register
       SPL      PLOW
       MPYK     1               ;Save T register (PR=TR)
       SPL      TREG
       SAR      6,Idc

;---------- Compute SPd = sum of Vd(k)*Idc(k)
       LT       Vd
       MPY      Idc
       PAC
       SACH     TEMP,I          ;Now this LS of 1 is for product mode only!
       ZALH     SPdH
       ADDS  SPdL
       ADD      TEMP
       SACL     SPdL            ;SPdL and SPdH are in DATA PAGE 6
       SACH     SPdH

;      ---------- CONTEXT RESTORE
CHEXT: ZALS     SACCL
       ADDH     SACCH           ;Restore ACC
       SPM      0
       LT       PLOW            ;Restore low P register
       MPYK     1               ;(TR)--->PRL
       LT       TREG            ;Restore T reg.
       LPH      PHIGH           ;Restore High P reg.
       LDPK     0
       LST1     SSV1            ;Restore status regs.
       LST      SSV0
```

```
        EINT                    ;Enable interrupts
        RET
; ---------------------------------------------------------------
;                AVERAGE COMPUTATION
; ---------------------------------------------------------------
; AVERAGE OF SELECTED VARIABLES OVER 1024 SAMPLES
AVE16:  LARP    5               ;Select AR5, down counter of 16
        BANZ    NZERO1          ;AR5=AR5-1,  and test if AR5>0
        LALK    1024
        SACL    TEMP0
        LAR     AR5,TEMP0       ;Case AR5 = 0; Reinitialize to 1024
        ZALH    SWeH            ; and get average of We
        ADDS    SWeL
;       SACH    Iqave,6         ;Iqave, average over 1024 samples (1.024 s)

        ZALH    SPdH2           ; Also get average of Pdc
        ADDS    SPdL2
        SACH    Pd16,6          ;Pd16, average over 1024 samples (1.024 s)
        ZAC                     ;Reset summation.
        SACL    SWeH
        SACL    SWeL
        SACL    SPdH2           ;Reset summation
        SACL    SPdL2

; ---------- Case AR5>0: Accumulate Iqs (Iqe) , Pdc (Ptemp)
NZERO1: ZALH    SWeH            ;Uses double precision to compute average.
        ADDS    SWeL
        ADD     Iqe
        SACL    SWeL
        SACH    SWeH

        LALK    14062           ;Convert Pdc1 into Watts
        SACL    PTEMP
        LT      Pdc1
        MPY     PTEMP
        PAC
        SACH    PTEMP           ;New Pdc in WATTS.
        ZALH    SPdH2           ;Uses double precision to compute average.
        ADDS    SPdL2
        ADD     PTEMP           ;Add Pdc1(WATTS) to summation, SPd (H,L).
        SACL    SPdL2
        SACH    SPdH2
        RET


; ---------------------------------------------
;   BUFFER STORAGE SUBROUTINE
; ---------------------------------------------
; ---------- DATA STORAGE CH-A
BUFSTR: ZALS    BUFAL
        ADDH    BUFAH
        ADDK    2
        SACH    BUFAH
        SACL    BUFAL
        SACH    TEMP0,7
        ANDK    1FEH
        SACL    LOW
        LAC     TEMP0
        ANDK    1FFH
        SACL    COL

        OUT     ONE,11
```

```
          IN        TEMP0,8
          OUT       LOW,ADDR
          OUT       COL,ADDR
          OUT       BFADATA,DATA   ;White line  in GRAPH
          OUT       ZERO,11

; ----------DATA STORAGE CH-B
          LALK      1FFH
          SUB       COL
          BGZ       JMPFUL
          LALK      0FE00H
          SACL      BUFAL
          LACK      3
          SACL      BUFAH
JMPFUL:   LAC       LOW
          ADDK      1
          SACL      LOW

          OUT       ONE,11
          IN        TEMP0,8
          OUT       LOW,ADDR
          OUT       COL,ADDR
          OUT       BFBDATA,DATA   ;Red line in GRAPH
          OUT       ZERO,11
          RET


; ---------------------------------------------------------------
;    OUTPUT INTERRUPT SERVICE ROUTINE
; ---------------------------------------------------------------
; ---------- Channel A
AINT:     EQU       $
          SST       SSV0
          LDPK      6
          OUT       OUTBFA, DAC    ;Load DAC LATCH
          NOP
          SXF                      ;Set output channel flag
          LDPK      0
          LST       SSV0
          EINT                     ;Enable interrupts
          RET                      ;Return from interrupt
;
; ---------- Channel B
BINT:     EQU       $
          SST       SSV0
          LDPK      6
          OUT       OUTBFB, DAC
          NOP
          RXF                      ;Reset output channel flag
          LDPK      0
          LST       SSV0
          EINT                     ;Enable interrupts
          RET                      ;Return from interrupt


; ---------------------------------------------------------------------------
;    FUZZY EFFICIENCY OPTIMIZATION ROUTINE - FEOPT
; ---------------------------------------------------------------------------
; ---------- START FEOPT
FEOPT:    SOVM                     ;Use overflow mode troughout FEOPT
          SPM       1
```

```
; ---------- Move some data from dp6 to dp7
          LDPK    6
          LAR     AR1,Wr
          LAR     AR2,Pdc
          LAR     AR3,TLest
          LDPK    7
          SAR     AR1,Wr2
          SAR     AR2,Pik
          SAR     AR3,TLest2

; ---------- Compute GID, the scale factor for DIds
          LT      TLest2
          ZALH    CG3             ;ACCH = C3
          MPY     CG2             ;P = C2*TLest
          LTA     CG1
          MPY     Wr2             ;ACCH = C2*TLest+C3
          APAC                    ;ACCH = C1*Wr+C2*TLest+C3 = GIDA
; ---------- Perform test for max |GID|
          ADDH    GIDMAX          ;Use saturation features to find max |GID|
          SUBH    GIDMAX
          SUBH    GIDMAX
          ADDH GIDMAX

; ---------- Perform test for min |GID|
          BGEZ    GGEZ            ;ACCH = GID
          ADDH  GIDMIN            ;Case GID < 0
          ADDH  GIDMIN
          SUBH    GIDMIN          ;Use saturation features to find min |GID|
          SUBH    GIDMIN          ;Actually GIDmin=(32768+minGID)/2
          B       GIDEND
GGEZ:  SUBH  GIDMIN               ;Case GID >= 0
          SUBH    GIDMIN
          ADDH  GIDMIN
          ADDH  GIDMIN
GIDEND: SACH    GId               ;Final value for GId
;
; ---------- Compute scale factor for DPi (GPb)
          ZALH    G2              ;ACCH = G2, DP = 7
          LT      G1
          MPY     Wr2
          APAC                    ;ACCH = G1*Wr + G2
; ---------- Test for max and min GPb
          ADDH  GPbmax    ;Max
          SUBH    GPbmax
          SUBH    GPbmin  ;Min
          SUBH    GPbmin
          ADDH  GPbmin
          ADDH  GPbmin
          SACH    GPb             ;Final value for  GPb

; ---------- Compute DPi (Pdc(k)-Pdc(k-1))*GPb
          ZALH    Pik
          SUBH    Pik1            ;ACCH = Pi(k)-Pi(k-1)
          ADDH  DPimax            ;Clamp |DPi| to DPimax
          SUBH    DPimax
          SUBH    DPimax          ;DPimax = 32767 - max DPi
          ADDH  DPimax

          SACH    TEMPi,7         ;Now TEMPi = (Pi(k)-Pi(k-1)) 2**7
          LT      GPb
          MPY     TEMPi           ;P=(Pi(k)-Pi(k-1)) 2**7 GPb
          PAC
```

```
          SACH    DPi             ;Final value for DPi
          DMOV    Pik             ;Pi(k)--> Pi(k-1)

; ---------- Compute Degree of Membership for DPi
          LAC     DPi
          BGEZ    DPiGEZ
          ADLK    P3              ;------- Case DPi<0
          SACL    TEMPi           ;Check if it is in the interval 0, -P3
          BLZ     NP1
          LT      RP3             ;Case -P3<DPi<0
          LACK    4               ;Interval index J = 4
          SACL    JP
          B       DPdone

NP1:      LAC     DPi             ;Compare DPi with -P2
          ADLK    P2
          BLZ     NP2
          LT      RP23            ;Case -P2 < DPi < -P3
          SACL    TEMPi
          LACK    3               ;Interval index J = 3
          SACL    JP
          B       DPdone

NP2:      LAC     DPi
          ADLK    P1
          LT      RP12            ;Case -P1<DPi<-P2
          SACL    TEMPi
          LACK    2               ;Interval index J = 2
          SACL    JP
          B       DPdone

DPiGEZ:   SBLK    P2              ;------- Case DPi > 0
          BLZ     POS1            ;Test if P2 < DPi < P1
          SACL    TEMPi           ;Case P2 < DPi < P1
          LACK    7
          SACL    JP              ;Interval index J = 7
          LT      RP12
          B       DPdone

POS1:     LAC     DPi
          SBLK    P3
          BLZ     POS2            ;Test if P3 < DPi < P2
          SACL    TEMPi           ;Yes
          LACK    6
          SACL    JP              ;Interval index J = 6
          LT      RP23
          B       DPdone

POS2:     LACK    5               ;Case  0 < DPi <P3
          SACL    JP              ;Interval index J = 6
          LT      RP3
          LAC     DPi
          SACL    TEMPi

DPdone:   MPY     TEMPi
          PAC                     ;ACC = MP2, D.M. of right MF for DPi
          SACH    MP2,2
          LAC     DMONE
          SUB     MP2             ;ACC = 1 - MP2 = MP1
          SACL    MP1
;
; ---------- Compute DM for LDIds (=DIds, up to this point)
```

```
        ZALH    DIds
        ADDH    L2lim           ;Limit |LDIds| to L2
        SUBH    L2lim
        SUBH    L2lim
        ADDH    L2lim
        SACH    DIds            ;Now |DIds|<= L2

        LAC     DIds            ;Get DM for Positive MF
        ADLK    L3
        SACL    TEMPi           ;ACC = LDIds + L3
        LT      RL23
        MPY     TEMPi
        PAC
        SACH    MI1             ;MI1 is the Positive MF of LDIds
        BGEZ    NEGMF
        ZAC
        SACL    MI1             ;MI1 = 0, since DPi < -L3

NEGMF:  ZAC                     ;Get DM for Negative MF
        SUB     DIds            ;ACC = -DIds
        ADLK  L3
        SACL    TEMPi           ;ACC = -DIds+L3
        MPY     TEMPi           ;T = RL23, unchanged
        PAC
        SACH    MI2
        BGEZ    IDONE           ;If MI2 > 0, the value is valid
        ZAC
        SACL    MI2             ;MI2 = 0, since computed DM<0

IDONE:  EQU     $

; ---------- Perform rule base evaluation (Min operator)
        LAC     MP1             ;MRA = MIN(MP1,MI1)
        SACL    MRA             ;Assume MRA = MP1
        SUB     MI1             ;ACC = MP1-MI1
        BLEZ    NXT1            ;Assumption is correct if ACC<=0
        LAC     MI1             ;No: MP1>MI1. Make MRA=MI1
        SACL    MRA

NXT1:   LAC     MP1             ;MRB = MIN(MP1,MI2)
        SACL    MRB             ;Assume MRB = MP1
        SUB     MI2             ;ACC = MP1-MI2
        BLEZ    NXT2            ;Assumption is correct if ACC<=0
        LAC     MI2             ;No: MP1>MI2. Make MRB=MI2
        SACL    MRB

NXT2:   LAC     MP2             ;MRC = MIN(MP2,MI1)
        SACL    MRC             ;Assume MRC = MP2
        SUB     MI1             ;ACC = MP2-MI1
        BLEZ    NXT3            ;Assumption is correct if ACC<=0
        LAC     MI1             ;No: MP2>MI1. Make MRC=MI1
        SACL    MRC

NXT3:   LAC     MP2             ;MRD = MIN(MP2,MI2)
        SACL    MRD             ;Assume MRD = MP2
        SUB     MI2             ;ACC = MP2-MI2
        BLEZ    NXT4            ;Assumption is correct if ACC<=0
        LAC     MI2             ;No: MP2>MI2. Make MRD=MI2
        SACL    MRD
NXT4:   EQU     $

; ---------- Get control signal for each of the 4 fired rules
```

```
; ---------- Rule A
          LAC     JP
          SUBK    2                 ;ACC = JP-2  (table offset)
          ADLK  EFTBL               ;Points to Eff. Optimization Table
          TBLR    DIA               ;Read DIA from table
          ADDK    1                 ;Change pointer to read DIC
          TBLR    DIC               ;ACC = JP-1
; ---------- Compute remaining control signals (DIB, DID)
          LALK    DMONE
          SUB     DIA
          SACL    DIB               ;DIB = 1 - DIA, for this RB
          LALK    DMONE
          SUB     DIC
          SACL    DID               ;DID = 1 - DIC, for this RB

; ------------------------------------------------------
; ---------- DEFUZZIFICATION BY HEIGHT METHOD
;
; ---------- Get sum of truth value of fired rules (MRA+MRB+MRC+MRD)
          LAC     MRA
          ADD     MRB
          ADD     MRC
          ADD     MRD
          SACL    SUMMR             ;SUMMR = MRA+MRB+MRC+MRD

;---------- Get sum of products: MRi*DIi, i=A,B,C,D
          LT      MRA
          MPY     DIA
          LTP     MRB               ;ACC = MRA*DIA
          MPY     DIB
          LTA     MRC
          MPY     DIC
          LTA     MRD
          MPY     DID
          APAC
          SACH    SUMPROD           ;SUMPROD=MRA*DIA+....+MRD*DID

; ---------- Divide SUMPROD BY SUMMR do get DIds (Fractional division)
;         Here, the sign of the result is defined by SUMPROD only
;         since SUMMR  is by def. always positive
          ZALH    SUMPROD
          ABS                       ;Make numerator positive
          RPTK    14
          SUBC    SUMMR  ;Result is in low ACC
          SACL    DIds
;
          LAC     SUMPROD
          BGEZ    DIVDONE           ;Done if sign is positive
          ZAC
          SUB     DIds
          SACL    DIds              ;Negate quotient if negative

; ---------- GET VALUE OF DIde IN AMPS
DIVDONE:  LT      GID
          MPY     DIds
          PAC
          SACH    TEMPi             ;TEMPi = DIde

; ---------- Impose mimimum value for DIde to keep search active
          ABS                       ;ACCH = |DIds|
          SUBH    DIdmin
          BGEZ    FEND
```

```
        LAC     TEMPi           ;Case |DIde| < DIdmin. Make DIde=DIdmin
        BGEZ    POSIde
        ZAC                     ;Case DIde < 0.
        SUB     DIdmin
        SACL    TEMPi           ;DIde = -DIdmin
        LALK    -L3
        SACL    DIds            ;Prevents |DIds| < L3
        B       FEND

POSIde: LAC   DIdmin            ;Case DIde >= 0
        SACL    TEMPi           ;DIde = +DIdmin
        LALK    L3
        SACL    DIds            ;Prevents |DIds| < L3

FEND:   ROVM
        LAC     TEMPi
        LDPK    6
        SACL    DIde            ;Store new DIde into dp 6
        RET
;
; ---------------------------------------------------
; ---------- DEFINE THE TABLE FOR DM's FOR DIds
EFTBL:  DW      I1,     I2,     I3,     0,      -I3,    -I2,    -I2


; ------------------------------------------------------------------
;       FEED FORWARD TORQUE COMPENSATOR
; ------------------------------------------------------------------
; ---------- NEW VALUE FOR Ide = AIde (Preliminary value)
FFTC: LDPK    6   ;Select DP 6
        SPM     1               ;For product shift correction
        LAC     Ide
        ADD     DIde            ;DIde from FEOPT
        SACL    AIde            ;Preliminary value of new Ide
;
; ---------- CHECK FOR MAX AND MIN Ide (SAFETY REASONS)
        SOVM                    ;Use saturation features for test purpose
        ZALH    AIde            ;Idmax=15729 (== 12 A), Idmin=3932 (== 3 A)
        ADDH    Idmax           ;Idmax = 32767-max Id
        SUBH    Idmax           ;End of MAX test

        SUBH    Idmin           ;Test for MIN Ide
        SUBH    Idmin           ;Subtract and add twice since
        ADDH    Idmin           ; ... Idmin = 0.5*(32768+Idmin)
        ADDH    Idmin           ;End of MIN test
        SACH    AIde            ;AIde now is between Idmin and Idmax
        SUBH    Ide             ;Corrects DIde in case AIde violates limits
        SACH    DIde            ;DIde = AIde-Ide

; ---------- COMPUTE ESTIMATE OF Lm
        LAC     AIde
        ADLK    -Idec0          ;Idec0 is the linear region breakpoint
        BGZ     GTRL            ;
        LALK    Lm0c            ;Case Ide<Idec0, linear region
        SACL    Lm              ;Lm = Lm0
        B       ENDL
GTRL:   SACL    TEMP             ;Case Ide>= Idec0, saturation region
        ZALH    Lm0             ;ACCH = Lm0
        LT      Mbs2
        MPY     TEMP            ;TEMP=Ide-Idec0
        SPAC                    ;ACC = Lm0-Mbs2*(Ide-Idec0)
        SACH    Lm
```

```
; ---------- ESTIMATE NEW VALUE OF ROTOR FLUX
ENDL:   LT      Lm
        MPY     AIde
        PAC                     ;ACC = Lm*AIde
        ABS                     ;Flux has to be always positive (Safeguard for div.)
        SACH    Nfdre           ;New steady state value of flux

; ---------- COMPUTE DIqe - STEP SIZE FOR Iqe
;          IMPLEMETATION: DIqe = 2 * {(fdre/2)/Nfdre - 1/2)} * Iqe
; ----------DIVIDE NUMERATOR (NUMKSL=Rr/Lr) BY Idef TO GET Kslip
        ZALH    fdre            ;Numerator is placed in High ACC
        RC                      ;0 --> C , Carry bit in ST1, before Rotation
        ROR                     ;Devide numerator by 2 (fdre/2)
; ---------- START FRACTIONAL DIVISION ROUTINE (Positive numbers only)
        RPTK    14
        SUBC    Nfdre           ;ACC = (fdre/2) / Nfdre
        ADLK    -16384          ;ACC = (fdre/2) / Nfdre  - 1/2
        SACL    TEMP,1          ;TEMP = 2 * {(fdre/2) / Nfdre - 1/2 }, 1 LSHIFT = x 2
        LT      TEMP
        MPY     Iqave
        PAC
        SACH    DIqe            ;

; ---------- GET NEW Iqe AND TEST FOR MAX STATOR CURRENT
        ADDH Iqe                ;ACC=Iqe+DIqe
        SACH    AIqe
        SQRA    AIqe            ;Get AIqe**2
        ZAC
        SQRA    AIde            ;Get AIde**2; ACC=AIqe**2
        APAC                    ;ACC = AIde**2 + AIqe**2 = Is**2
        SACH    IP2             ;IP2 = Is**2
        LAC     IP2
        ADLK  -IMAX             ;IMAX is the square of max. stator current
        BGZ     GTRIS           ;Branch if Is**2 > IMAX
        LAC     SDIqe           ;Case Is**2 < IMAX
        ADD     DIqe            ; Update SDIqe, sum of DIqe
        SACL    SDIqe
        DMOV  AIde              ;AIde --> Ide
        DMOV  Nfdre             ;Nfdre --> fdre for next iteration
        B       ENDIS

GTRIS:  ZAC                     ;Case Is**2 > IMAX, neglect changes
        SACL    DIqe

ENDIS:  SPM     0               ;Restore for compatibility with old routines
        RET


; ---------------------------------------------------------
;     FIRST ORDER FILTER FOR Pdc SIGNAL
; ---------------------------------------------------------
;          TAL = 500 ms
;          Tsamp = I ms
; ************************************        ****************************
; *        y(k) = c x(k) + d u(k)        *        * y<==>Pdc    u<==>PdcI      *
; *        x(k+I) = a x(k) + b u(k)      *        * a<==>A3     b<==>B3        *
; ************************************        * c<==>C3     d<==>D3        *
;                                             * x(k)<==>xpk x(k+1)<==>xpk1*
;                                             ****************************
;
; ---------- COMPUTE Pdc - Filtered version of PdcI
FOFP:   LDPK    6
```

```
        SPM     1
        ZAC
        LT      xpk
        MPY     C3
        LTA     Pdc1            ;ACC = c x(k) , T = u(k)
        MPY     D3
        APAC                    ;ACC = c x(k) + d u(k)
        SACH    Pdc
        ZAC
        MPY     B3              ;T = u(k) has not changed
        LTA     xpk             ;ACC = b u(k) , T = x(k)
        MPY     A3
        APAC                    ;ACC = a x(k) + b u(k)
        ADDS    XPL             ;Add 16 lsb from last iteration
        SACH    xpk             ;Actually x(k+1) ---> x(k) for next iteration
        ANDK    65280           ;Get upper half of ACCL
        SACL    XPL             ;Save 16 lsb for next iteration
        RET


; ------------------------------------------------------------------------------
;   SLIP GAIN COMPUTATION, INCLUDING FIRST ORDER FILTER
; ------------------------------------------------------------------------------
;       Tr = 0.1073s (Rotor time constant)
;       Tsamp = 1 ms
; ************************          ************************
; *     y(k) = c x(k) + d u(k)*     * y<==>Idef   u<==>Nfdre     *
; *     x(k+1) = a x(k) + b u(k)*   * a<==>A1     b = 1          *
; ************************          * c<==>C1     d<==>D1        *
;                                   * x(k)<==>xks x(k+1)<==>xk1s *
;                                   ************************
FOF1:   SPM     1
        ZAC
        LT      xks
        MPY     C1
        LTA     Nfdre           ;ACC = c x(k) , T = u(k)
        MPY     D1
        APAC                    ;ACC = c x(k) + d u(k)
        SACH    Idef
        ZAC
        MPY     B1              ;T = u(k) has not changed
        LTA     xks             ;ACC = b u(k) , T = x(k)
        MPY     A1
        APAC                    ;ACC = a x(k) + b u(k)

        ADDS    XKLS            ;Add 16 lsb from last iteration
        SACH    xks             ;Actually x(k+1) ---> x(k) for next iteration
        ANDK    65280           ;Take upper half of ACCL (65280=FF00h)
        SACL    XKLS            ;Save 16 lsb for next iteration

; ----------DIVIDE NUMERATOR (NUMKSL=Rr/Lr) BY IdefTO GET Kslip
        LAC     Idef
        ABS                     ;Get abs of Idef to avoid error (Idef>0 always)
        SACL    Idef            ;
        ZALH    NUMKSL          ;Numerator is placed in High ACC

; ---------- START DIVISION ROUTINE (Positive numbers only)
        RPTK    14
        SUBC    Idef
        SACL    Kslip
        RET
```

```
; ------------------------------------------------------------------------------
;       FIRST ORDER FILTER FOR FFTC - COMPUTATION OF SDIqf
; ------------------------------------------------------------------------------
;           Tr = 0.1819s (Rotor time constant)
;           Tsamp = 1 ms
; ***************************           *******************************
; *     y(k) = c x(k) + d u(k)    *       * y<==>Idef   u<==>SDIqe      *
; *     x(k+1) = a x(k) + b u(k)  *       * a<==>A1     b = B1          *
; ***************************           * c<==>C1     d<==>D1          *
;                                       * x(k)<==>xkf x(k+1)<==>xk1f *
;                                       *******************************
; ----------- COMPUTE SDIqf - Filtered version of SDIqe
FOFT:   LDPK    6
        SPM     1
        ZAC
        LT      xkf
        MPY     C1
        LTA     SDIqe           ;ACC = c x(k) , T = u(k)
        MPY     D1
        APAC                    ;ACC = c x(k) + d u(k)
        SACH    SDIqf
        ZAC
        MPY     B1              ;T = u(k) has not changed
        LTA     xkf             ;ACC = b u(k) , T = x(k)
        MPY     A1
        APAC                    ;ACC = a x(k) + b u(k)

        ADDS    XKLT            ;Add 16 lsb from last iteration
        SACH    xkf             ;Actually x(k+1) ---> x(k) for next iteration
        ANDK    65280           ;Take upper half of ACCL (65280=ff00h)
        SACL    XKLT            ;Save 16 lsb for next iteration

; ----------- GET COMPENSATED Iqe
        LAC     Iqep            ;Iqep from PI speed controller
        ADD     SDIqf           ;Iqe = Iqep + SDIqf
        SACL    Iqe

; ----------- COMPUTE TORQUE ESTIMATE, TLest
        LT      Iqave           ; As FEOPT is called before FFTC, we use
        MPY     fdre            ; Iqave(Iqs*) and fdre to estimate TL
        PAC
        SACH    TLest
        RET


; ------------------------------------------------------------------------------
;     FIRST ORDER FILTER FOR Iqs* (Used in FFTC only)
; ------------------------------------------------------------------------------
;           TAL = 0.1 sec
;            Tsamp = 1 ms
; ************************************          ****************************
; *       y(k) = c x(k) + d u(k)         *       * y<==>Iqave  u<==>Iqe      *
; *       x(k+1) = a x(k) + b u(k)       *       * a<==>A2     b<==>B2      *
; ************************************          * c<==>C2     d<==>D2      *
;                                               * x(k)<==>xwk x(k+1)<==>xwk1*
;                                               ****************************
; ----------- COMPUTE Iqave, filtered version of Iqs*
FOFW:   LDPK    6
        SPM     1
        ZAC
```

```
        LT      xwk
        MPY     C2
        LTA     Iqe             ;ACC = c x(k) , T = u(k)
        MPY     D2
        APAC                    ;ACC = c x(k) + d u(k)
        SACH    Iqave           ;Correct result before storing ( * 2 )
        ZAC
        MPY     B2              ;T = u(k) has not changed
        LTA     xwk             ;ACC = b u(k) , T = x(k)
        MPY     A2
        APAC                    ;ACC = a x(k) + b u(k)
        ADDS    XKL             ;Add 16 lsb from last iteration
        SACH    xwk             ;Actually x(k+I) ---> x(k) for next iteration
        ANDK    65280           ;Get upper half of ACCL
        SACL    XKL             ;Save 16 lsb for next iteration
        RET

; ------------------------------------------------------------
;    HOST COMMUNICATION ROUTINE
; ------------------------------------------------------------
        LAC     MSTEP
        BZ      SBF0            ;MSTEP=0
        SUBK    I
        BZ      SBF1            ;MSTEP=1
        SUBK    1
        BZ      SBF2            ;MSTEP=2
        ZAC                     ;MSTEP value is incorrect, make MSTEP=0
        SACL    MSTEP

; ---------- CASE MSTEP=0 : READ FIRST VALUE FROM HOST (Code)
SBF0:   IN      HOSTBUF, HOSTSTAT;Read host status
        DINT
        BIT     HOSTBUF, 15     ;Test host status
        BBZ     MAIN            ;Loop if no data is available
        EINT
        IN      COMMAND, READHOST       ;Read host 1st data (COMMAND)
        CALL    CASEA                   ;Single command functions
        B       MAIN

; ---------- CASE MSTEP=1 : READ SECOND VALUE FROM HOST (Par. value)
SBF1:   IN      HOSTBUF, HOSTSTAT       ;Read host status
        DINT
        BIT     HOSTBUF, 15             ;Test host status
        BBZ     MAIN                    ;Loop if no data is avaiable
        EINT
        IN      COMMAND2, READHOST  ;Read host 2nd data (COMMAND2)
        LACK    2                       ;Make MSTEP=2
        SACL    MSTEP
        B       MAIN

; ---------- CASE MSTEP=2 : FUNCTIONS THAT REQUIRE 2 COMMANDS.
;          IDENTIFY AND STORE INFORMATION IN PROPER PARAM./VAR.
SBF2:   ZAC                             ;Set MSTEP=0
        SACL    MSTEP

; ---------- S0 THROUGH S6 TEST FOR EACH PARTICULAR COMMAND(INSTRUCTION)
S0:     LAC     COMMAND
        SUBK    2                       ;COMMAND=2? (Wref input)
        BNZ     S1                      ;NO: Test next condition

        SPM     2
        ZAC
```

```
        LT      COMMAND2        ;YES: Read Wref (in rpm)
        MPY     KWREF           ; and convert it to software value
        APAC
        SACH    WREF            ;Now Wref is correct software value
        SPM     0
        B       CMDEND

S1:     LAC     COMMAND
        SUBK    6
        BNZ     S2              ;COMMAND=6 (Def. Buffer A var.)
        LAC     COMMAND2
        SACL    VAR1            ;VAR1 is the code for 1st var.
        B       CMDEND

S2:     LAC     COMMAND
        SUBK    7               ;COMMAND=7 (Def. Buffer B var.)
        BNZ     S3
        LAC     COMMAND2
        SACL    VAR2            ;VAR2 is the code for 2nd var.
        B       CMDEND

S3:     LAC     COMMAND
        SUBK    8
        BNZ     S4              ;COMMAND=8 (Input Poffset)
        LAC     COMMAND2        ; Offset value for Pdc.
        SACL    Poffset
        B       CMDEND

S4:     LAC     COMMAND
        SUBK    9               ;COMMAND=9 (Input K1 of PI
        BNZ     S5              ;  speed controller)
        LAC     COMMAND2
        SACL    K1
        LDPK    6
        B       CMDEND

S5:     LAC     COMMAND
        SUBK    10              ;COMMAND=10 (Input K2 of PI
        BNZ     S6              ;  speed controller)
        LAC     COMMAND2
        SACL    K2
        B       CMDEND

S6:     LAC     COMMAND         ;DEBUG FUNCTION
        SUBK    11              ;COMMAND=11 (Input Ide (ids*)
        BNZ     CMDEND          ; set S/W timer 3 to 2500 and
        LAC     COMMAND2        ; reset buffer storage
        SACL    Ide
        LALK    10000
        SACL    SWT3
        B       CMDEND

CMDEND: ZAC                     ;If the instruction was acknowledged
        SACL    MSTEP           ; set MSTEP=0 and loop again
        B       MAIN

; ---------- RESET BUFFER STORAGE SUBROUTINE
BUFSTO: DINT
        ZAC
        SACL    BUFAH
        SACL    BUFAL
        SACL    COL
```

```
        SACL    LOW
        EINT
        RET

; ---------- CASE A: Single instruction routines

CASEA:  LAC     COMMAND             ;COMMAND=0 :Stop system
        BNZ     CMD1
        LACK    1
        SACL    INCOMDATA
        OUT     INCOMDATA,INCOM     ;INCON=1: Disable inverter gate
        B       AEND

CMD1:   LAC     COMMAND             ;COMMAND=1 : Set rated flux
        SUBK    1
        BNZ     CMD2
        LACK    3
        SACL    INCOMDATA
        OUT     INCOMDATA,INCOM     ;INCON=3 : Enable inverter gate
        B       AEND

CMD2:   LAC     COMMAND             ;COMMAND=3: Start Eff. Opt. mode
        SUBK    3                   ; with Buffer storage
        BNZ     CMD3
        LACK    1
        SACL    EOM                 ;Set Efficiency Optimiz. Mode (=1)
        CALL    BUFSTO
        LALK    10000               ;Load S/W TIMER 3 for 0.5 s period
        SACL    SWT3
        B       AEND

CMD3:   LAC     COMMAND             ;COMMAND=4: Enter dynamic mode
        SUBK    4
        BNZ     CMD4
        ZAC
        SACL    EOM                 ;Reset Efficiency Optimiz. Mode (=0)
        B       AEND

CMD4:   LAC     COMMAND             ;COMMAND=5: Start buffer storage
        SUBK    5                   ;  VAR1 and VAR2 define the vars.
        BNZ     CMD5                ; being stored
        CALL    BUFSTO
        B       AEND

CMD5:   LALK    1                   ;Case of 2 instr. commands
        SACL    MSTEP               ; Make MSTEP=1 and return
        RET

AEND:   ZAC                         ;If the instruction was acknowledged
        SACL    MSTEP               ; set MSTEP=0  and return
        RET

; ------------------------------------------------------------------------
;     ROUTINE FOR Pdc (average) COMPUTATION
; ------------------------------------------------------------------------
; NOTES: 1) AVERAGE IS COMPUTED EVERY 200 ms INTERVAL;
;        2) SPd=(SPH,SPdL) IS THE SUM OF Vd(k)*Idc(k) OVER 200 ms
;           (TRAPEZOIDAL RULE OF INTEGRATION), FROM ADC ISR
;        3) THE AVERAGE IS COMPUTED BY MAKING
;           SUM OF Vd(k)*Idc(k) FOR k = 1,10000} * {1/10000}
;        4) TO INCREASE RESOLUTION, THE SUM IS FIRST MULTIPLIED BY 2^3
;           AND THEN  DIVIDED BY 2**16  (TAKING THE ACCH)
```

```
;          5) IN ORDER TO HAVE CORRECT SCALE, THE RESULT IS
;             MULTIPLIED BY 26844 (= 1/1.2207).OVERALL EFFECT IS:
;             (1/10000) = 2**3 / {2**16 * 1.2207)
;
;
PDCAVE:          SPM     1
          LDPK   6
          ZALH   SPdH              ;Store full SPd into ACC
          ADDS   SPdL
          ABS                      ;Only positive Pdc are possible with diode rect.
          SACH   Pdc1,3            ;This value is actually 1.2207 * (true Pdc)
          ZAC
          SACL   SPdH              ;Reset SPd for new 50 ms interval
          SACL   SPdL

; ---------- COMPUTE TRUE Pdc
          LALK   17619             ;Constant to convert Pdc to its
          SACL   PTEMP             ;correct scaled value (k/2 = 17619)
          LT     Pdc1
          MPY    PTEMP
          PAC
          SACH   Pdc1,1            ;Left shift 1 to get proper scale
          RET

; ---------------------------------------------------------------------------
;          PI SPEED CONTROLLER FOR VECTOR CONTROL SYSTEM
; ---------------------------------------------------------------------------
 PI:      LDPK   6                 ;USE PAGE 6
          SPM    1
          LAC    WREF              ;ACC = WREF
          SUB    WR                ;ACC = WREF-WR
          SACL   EK                ;e(k) = WREF-WR
          ZALH   IQH               ;ACC = iqs*(k-1)
          ADDS   IQL
          LT     EK1               ;T = e(k-1)
          MPY    K2                ;P = K2*e(k-1)
          LTD    EK                ;e(k)->T, e(k)-->e(k-1), iqs*(k-1)+K2*e(k-1)-->ACC
          MPY    K1
          APAC                     ;ACC = iqs*(k-1)+K2*e(k-1)+K1*e(k) = v
          SACH   IQH
          SACL   IQL

; TEST FOR MAX AND MIN iqs* (u = f(v))
          SOVM                     ;Set overflow mode
          ADDH   ILIM              ;Test for max iqs*
          SUBH   ILIM
          SUBH   ILIM              ;Test for minimum iqs*
          ADDH   ILIM
          SACH   Iqep,6            ;Iqep (mult. by 64 = 2**6 for better precision)
          ROVM                     ;To prevent error in other routines
          RET


; ---------------------------------------------------
;     SPEED FEEDBACK SUBROUTINE
; ---------------------------------------------------
; NOTE:  Modified for  5 hp Class B machine.
;                (Pulse encoder: 14400 pulses/rev.)
SPEED:  LDPK     6
          SPM    0
          LAC    SSAMP             ;Increment iteration counter SSAMP
          ADDK   1
```

```
        SACL    SSAMP
        LAC     STP             ;STP = 0 --> Read M, T counts
        BZ      STP0
        SUBK    1
        BZ      STP1            ;STP = 1 --> Compute new Wr
        ZAC
        SACL    STP             ;If STP not 0 or 1, reset to 0

STP0:                           ;  STEP 0 : Read M (P.G.) and T (ClocK)
        LAC     SSAMP
        SBLK    SSAMPC          ; =1msec
        BLZ     SFEXIT

        IN      MNEW,MCNT       ;Input M, T counters content
        IN      TNEW,TCNT
        IN      TEMP0,MCNT      ;Read MCNT again for security reasons
        LAC     TEMP0
        SUB     MNEW
        BNZ     SFEXIT          ;If TEMP0 not = MNEW, neglect readings
        LAC     MNEW            ; New data has been latched?
        SUB     MOLD
        BZ      STP02           ; No: Check if max Ts has elapsed.

; ---------- Case MNEW not= MOLD
        LAC     SSAMP           ; Greater than max sampling time?
        SBLK    STMAXC
        BGZ     STP01           ; Yes: speed is close to 0. Branch to STP01
;                               ; No: Data is valid; proceed.
        LAC     MNEW            ; calc. delta M
        SUB     MOLD
        SACL    DELM,6          ; dM=32*(MNEW-MOLD ) FOR CLASS B
        LALK    18641                   ; USE NEW PULSE ENCODER
        SACL    TEMP0           ; FACTOR = 0.5689
        LT      TEMP0           ; Overall: 1/7.0313 = 1/4 * 0.5689
        MPY     DELM
        PAC
        SACH    DELM,1
        LAC     TNEW
        SUB     TOLD
        SACL    DELT            ;DELT = TNEW - TOLD
        LACK    1               ; Next iteration execute STP1
        SACL    STP
        B       SFEXIT

STP01:  ZAC                      ; SSAMP = 0, Wr = 0
        SACL    SSAMP
        SACL    DELM
        DMOV    MNEW            ; MOLD <-- MNEW
        DMOV    TNEW            ; TOLD <-- TNEW
        DMOV    Wr1
        B       SFEXIT

STP02:  LAC     SSAMP
        SBLK    STMAXC
        BLZ     SFEXIT
        LACK    STMAXC          ; SSAMP = Stmax, Wr = 0
        SACL    SSAMP
        ZAC
        SACL    DELM            ; Originally was Wr
        B       SFEXIT

STP1:   EQU     $               ; Compute new speed value
```

```
        LAC     DELM,15         ; WRTEMP = DELM / DELT
        ABS
        LARP    1
        LARK    1,15

STP11: SUBC     DELT
        BANZ    STP11

DTST: IN TEMP0,INSTU            ;Direction check.
        BIT     TEMP0,15
        BBZ     STP12
        NEG                     ;Negative speed.

; ---------- SPEED VAR. LIMIT
STP12:  SOVM                    ; Set overflow mode to use saturation feature

        SUB     WROLD
        SACL    TEMP0           ;TEMP0 = Wrnew - Wrold
        ZALH    TEMP0
        ADDH    DELWRC          ;Test for DELWRC > +DELWRC
        SUBH    DELWRC
        SUBH    DELWRC          ;Test for DELWRC < -DELWRC
        ADDH    DELWRC
        ADDH    WROLD
        SACH    Wr1             ;New value of Wr1, within limits of DELWRC
        ZAC                     ;Reset counters
        SACL    SSAMP
        SACL    STP
        DMOV    MNEW            ; MOLD <-- MNEW
        DMOV    TNEW            ; TOLD <-- TNEW
        DMOV    Wr1             ; WROLD <-- Wr

SFEXIT:  RET


; ---------------------------------------------------------------------------
;               TRANSITION CONTROL SUBROUTINE
; ---------------------------------------------------------------------------
TRCTL:  LAC   FSS
        BGZ     FSS1            ;FSS=1 --> Steady state mode.
FSS0:   LAC     EK              ;EK = Wref-Wr
        ABS
        SBLK    TOLst
        BLEZ    CSST1
        RET                     ;If EK> TOLst, stay in Dyn. mode

CSST1:  LAC  CSS                ;EK<TOLst
        ADDK  1                 ;Increment counter for steady state (ss) test
        SACL    CSS
        SUBK    3               ;CSS-3 >= 0 ?
        BGEZ    SS
        RET                     ;No: stay in dynamic mode

SS:     LACK  1                 ;Yes: ss mode has been detected.
        AND     EOM             ;AND with EOM flag, to allow operator control
        SACL    FSS             ;FSS = 1 indicates ss mode
        BZ      RTN             ;If FSS = 0, it is because EOM=0!

; ---------- Initialize Pik1 and DIds for the first iteration
        LDPK    6
        LALK    10000
        SACL    SWT3
```

```
        LALK    1400
        ADD     Pdc
        LDPK    7
        SACL    Pikl            ;Initialize Pikl to Pdc+1400, such that DPi=NB
        LALK    -32767
        SACL    DIds            ;Initialize LDIds to NB

RTN:    ZAC
        SACL    CSS             ;Reset counter
        RET

;--------- Case FSS = I. Check if a dyn. mode has come
FSS1:   LAC     EOM             ;Check if operator has terminated eff. opt. mode
        BZ      DYN             ;If EOM = 0, exit eff. opt mode, regardless of EK
        LAC     EK
        ABS
        SBLK    TOLex
        BGZ     CSST2           ;EK > TOLex ?
        RET                     ;No: stay in ss mode

CSST2:  LAC  CSS                ;Yes: Make futher tests.
        ADDK    1               ;Increment counter, for dynamic test
        SACL    CSS
        SUBK    3               ;CSS >= 3  ?
        BGEZ    DYN
        RET                     ;No: stay in ss mode

DYN:    ZAC                     ;Yes: make transition to dynamic mode
        SACL    FSS             ;FSS = 0 indicates ss mode
        SACL    CSS             ;Reset counter
        LALK    Iderated        ;Restore rated flux
        SACL    Ide
        LALK    6335            ;Rated flux for 5 hp class B m/c.
        SACL    fdre            ;Reset flux to rated value
        RET


; ------------------------------------------------------------------------
;       PERFORM BASIC VECTOR CONTROL FUNCTIONS
; ------------------------------------------------------------------------
VECROT: SPM     0
        LDPK    6
        ZALH    Wr              ;We computation. We = Wr + Kslip*Iqe
        LT      Kslip           ;As PM=0, net effect is Wr+(Kslip/2)*Iqe
        MPY     Iqe
        APAC
        APAC
        SACH    We

; ---------- Calc. angle theta
BP1:    ROVM                    ;Reset overflow mode
        ZALH    THETAH
        ADDS    THETAL
        LT      We
        MPYK    DT
        APAC
        SACL    THETAL
        SACH    THETAH
;
;---------- Get  SIN(theta), COS(theta) from look-up table
        SSXM                    ;Set sign extension
        LT      THETAH
```

```
        MPYK    2048
        PAC
        SACH    TEMP0
        LAC     TEMP0
        BGEZ    POSIN
        ADLK    1024            ;If negative, add 180 degrees
        ANDK    1023
;
NESIN:  SOVM
        ADLK    SINTBL
        TBLR    SIN
        ADLK    TBL90
        TBLR    COS
        LAC     SIN
        NEG
        SACL    SIN
        LAC     COS
        NEG
        SACL    COS
        B       SINEND

POSIN:  SOVM                    ;Set overflow mode
        ADLK    SINTBL
        TBLR    SIN
        ADLK    TBL90
        TBLR    COS
SINEND: EQU     $

; ---------- Syncrhonous to stationary conversion
;       Iqs= Iqe*COS+Ide*SIN
;       Ids=-Iqe*SIN+Ide*COS
        ZAC
        LT      SIN
        MPY     Ide
        LTA     COS
        MPY     Iqe
        APAC
        SACH    Iqs,1
        ZAC
        MPY     Ide
        LTA     SIN
        MPY     Iqe
        SPAC
        SACH    Ids,1
;
; ---------- 2 TO 3 Phase conversion
        LAC     Iqs
        SACL    Ias
        ZAC
        SUB     Iqs,14
        LT      Ids
        MPY     RT34
        SPAC
        SACH    Ibs,1
        LAC     Ibs
        RET             ; ---------- End of vector rotation
```

B.2 Slip Gain Tuning Assembly Programs.

```
; ------------------------------------------------------------------
;       FUZZY LOGIC BASED SLIP GAIN TUNING
;         FOR VECTOR CONTROL AC200 SYSTEM
; ------------------------------------------------------------------
;
;  GILBERTO C. D. SOUSA
;  UNIVERSITY OF TENNESSEE, KNOXVILLE.
;
;THIS VERSION INCLUDES:
;           1 - VECTOR CONTROL FUNCTIONS
;           2 - SPEED COMPUTATION ROUTINE
;           3 - PI SPEED CONTROL
;           4 - ADC ISR FOR TWO CHANNELS
;           5 - DAC ISR
;           6 - REVERSE VECTOR ROTATION FOR Vds AND Vqs
;           7 - FILTER FOR Vds AND Vqs
;           8 - DELTA Q, DELTA Vd COMPUTATION
;           9 - RULE BASE I - Kf COMPUTATION
;           10 - RULE BASE II - DELTA Ks COMPUTATION

; ------------------------------------------------
;               I/O PORT ADDRESS
; ------------------------------------------------
MCNT:           EQU     0       ;pulse cnt. read
TCNT:           EQU     1       ;clock      cnt. read
INSTU:          EQU     2       ;inverter status READ
INCOM:          EQU     2       ;inverter command WRITE

DATA:           EQU     8       ;Buffer data port
ADDR:           EQU     9       ;Buffer address port
READHOST:       EQU     10      ;Host read port
REFRESH:        EQU     11      ;Enable buffer refresh
DAC:            EQU     13      ;DAC latch
HOSTSTAT:       EQU     14      ;Host status
WRITEHOST:      EQU     15      ;Host write port

; ------------------------------------------------
;                   CONSTANTS
; ------------------------------------------------
PRDC:           EQU     2000    ;MAIN SAMPLING TIME=.2msec
Iderated:       EQU     12080   ;RATED Ids = 9.216 A (= 12080)
SLIPGAIN:       EQU     244     ;UNDER RATED Ids, SLIPGAIN==244
DT:             EQU     4000    ; FOR THETA CALC.
TBL90:          EQU     512     ;90 DEGREE
SSAMPC:         EQU     5       ;SPEED SAMPLING=1msec
STMAXC:         EQU     10      ;15=3msec MAX. SPEED SAMP. TIME
DELWRCC:        EQU     32767-100 ; Max speed VARIATION IN 1 MS

; ---------- MEMORY MAPPED REGISTERS (ALREADY DEF. IN RESMON
;DRR:           EQU     0       ;A/D Input data register
;TIM:           EQU     2       ;TIMER
;PRD:           EQU     3       ;PERIOD REG.
;IMR:           EQU     4       ;Interrupt mask register
;GREG:          EQU     5       ;Global memory allocation register
;
; ---------- DEFINE CONSTANTS FOR PI CONTROLLER
K1C:            EQU     19637   ;For kp=2 and ki=0.2: K1C=19637
K2C:            EQU     -19635 ;                      : K2C=-19635
IQMV:           EQU     391     ;|Iqe| limit / 64 == 391
                                ;Maximum value of Iqs*=19.11 A == 25047
```

```
; ---------- DEFINE CONSTANTS FOR FIRST ORDER FILTER: Vds, Vqs.
;          CURRENTLY USING TAL = 2 ms, fsamp = 25 KHz.
A1C:            EQU     29647
B1C:            EQU     3121
C1C:            EQU     31208
D1C:            EQU     1560

; ---------- DEFINE CONSTANTS FOR FOF OF FEEDBACK SPEED
A2C:            EQU     31357
B2C:            EQU     2048
C2C:            EQU     22090
D2C:            EQU     706

; ---------- DEFINE CONSTANTS FOR FIRST ORDER FILTER: CE
;            CURRENTLY USING TAL = 30 ms, Ts = 1ms.
A3C:            EQU     31694
B3C:            EQU     1074
C3C:            EQU     32231
D3C:            EQU     537

; ---------- CONSTANTS FOR RULE BASE II, DKs COMPUTATION.
e1:             EQU     -19660  ;Membership functions for Error
e2:             EQU     -9830
e3:             EQU     -1966
e4:             EQU     3277
e5:             EQU     16384
e6:             EQU     32767

ce1:            EQU     6553    ;Membership function for CE.
ce2:            EQU     19660
ce3:            EQU     32767

dk1:            EQU     -32767  ;Rule base for DKs.
dk2:            EQU     -19660
dk3:            EQU     -13107
dk4:            EQU     -1311
dk5:            EQU     4915
dk6:            EQU     9830
dk7:            EQU     19660
dk8:            EQU     32767

; ------------------------------------------------------
;              DEFINE DATA RAM
; ------------------------------------------------------
;PAGE 0          SSV0 and SSV1 must be in PAGE 0!
                ORG     061H
SSV0:           DS      1
SSV1:           DS      1
SSTEMP:         DS      1
; ------------------------------------------------------

;PAGE 6
                ORG     00h     ;Start of on-board data RAM (PAGE6)
OUTBFA:         DS      1       ;Channel A output buffer
OUTBFB:         DS      1       ;Channel B output buffer
COMMAND:        DS      1       ;Host command word
COMMAND2:       DS      1       ;Value passed to TMSC25
DISP:           DS      1       ;Value passed to HOST
HOSTBUF:        DS      1       ;Buffer host status

;---------- MAIN INT.
THETAL:                 DS      1
```

```
THETAH:         DS      1
We:             DS      1
Kslip:          DS      1
SIN:            DS      1
COS:            DS      1
Iqs:            DS      1
Ids:            DS      1
Ias:            DS      I
Ibs:            DS      1
RT34:           DS      1

;---------- DATA SAVE
BUFAL:          DS      1
BUFAH:          DS      1
COL:            DS      1
LOW:            DS      1
BFADATA:        DS      1
BFBDATA:        DS      1
TEMP2:          DS      1
SACCH:          DS      1
SACCL:          DS      1
ZERO:           DS      1
ONE:            DS      1

; ---------- SPEED FEEDBACK
STP:            DS      1
SSAMP:          DS      1
MNEW:           DS      1
MOLD:           DS      1
TNEW:           DS      1
TOLD:           DS      1
Wr1:            DS      1
WROLD:          DS      1
DELM:           DS      1
DELT:           DS      1
DELWRC:         DS      1

; ---------- VECTOR CONTROL VARIABLES
Ide:            DS      1
Iqe:            DS      1
PHIGH:          DS      1
PLOW:           DS      1
TREG:           DS      1

;---------- DEFINE VARIABLES FOR PI CONTROLLER
IQH:            DS      1
IQL:            DS      1
ILIM:           DS      1       ;Limit for test of iqs*max (32767-IQMAX)
EK:             DS      1       ;Speed error e(k)
EK1:            DS      1       ;e(k-1)
WREF:           DS      1       ;Speed reference
KWREF:          DS      1       ;Constant for RPM-->decimal convertion
KBR:            DS      1       ;  "     " decimal -->RPM convertion
IQMAX:          DS      1       ;IQMAX location
K1:             DS      1       ;PI gains (derived from, but not equal, kp,ki).
K2:             DS      1

; ---------- VARIABLES FOR HOST/DSP COMMUNICATION
INCOMDATA:      DS      1
MSTEP:          DS      1       ;MAIN ROUT. STEP
VAR1:           DS      1       ;Buffer storage var. 1 (Ch. A)
VAR2:           DS      1       ;Buffer storage var. 2 (Ch. B)
```

```
; ---------- TIMERS FOR REAL TIME SCHEDULER
SWT0:           DS      1       ;S/W timer 0: variable Ts, for Buffer storage
SWT1:           DS      1       ;S/W timer 1: 1 ms for PI speed and SGT.
SWT2:           DS      1       ;Test of speed response

; ---------- VARIABLES FOR FEEDBACK SPEED SIGNAL
A2:             DS      1       ;Coefficients for filter FOFW
B2:             DS      1
C2:             DS      1
D2:             DS      1
XWK:            DS      1       ;State variable
Wr:             DS      1       ;Filtered speed signal
XKL:            DS      1       ;16 lsb of xkw          (Speed fb)

; ---------------------------------------------------------
; ---------- VARIABLES FOR SLIP GAIN ROUTINES
; PAGE 6, CONTINUED.

; ---------- VARIABLES FOR IVR.ASM: INVERSE VECTOR ROTATION
SVdH:           DS      1
SVdL:           DS      1
SVqH:           DS      1
SVqL:           DS      1
AV:             DS      1       ;Coefficients for temp. compensation
BVs:            DS      1       ; of Vds and Vqs sensing.
RT123:          DS      1       ;1 / sqrt(3).
Talv:           DS      1       ;Tal for phase shift correction.

FSG:            DS      1       ;Flag for slip gain process.
Vds :           DS      1       ;Stationary frame D-Q voltages.
Vqs:            DS      1
Vde:            DS      1       ;Synchronous frame D-Q voltages
Vqe:            DS      1
TEMP0:          DS      1       ;Auxiliary variable
SIND:           DS      1       ;Unit vector with phase shift compensation.
COSD:           DS      1

; ---------- VARIABLES FOR FILTER.ASM, Vde AND Vqe FILTER.
xkd1:           DS      1       ;Vde state variable for 1st filter.
xkd2:           DS      1       ;Vde state variable for 2nd filter.
Vde1:           DS      1       ;Output of first Vde filter.
Vdef:           DS      1       ;Final Vde, after 2nd filter.
xkq:            DS      1       ;Vqe state variable.
Vqef:           DS      1       ;Filtered Vqe.
xd1l:           DS      1       ;16 lsb of xkd1
xd2l:           DS      1       ;16 lsb of xkd2
xql:            DS      1       ;16 lsb of xkq
A1:             DS      1       ;Constants for filter Vde, Vqe filters
B1:             DS      1
C1:             DS      1
D1:             DS      1

; ---------- VARS. FOR DELTAC.ASM, DELTA Q AND DELTA Vd COMP.
QE:             DS      1       ;Actual Q.
QEC:            DS      1       ;Command Q.
Qb:             DS      1       ;Base Q (= Kq+QEC).
Vdec:           DS      1       ;Command Vd.
Vqec:           DS      1
Vb:             DS      1       ;Base Vd.
TEMP1:          DS      1       ;Auxiliary variable.
KQ:             DS      1       ;Epson for Qb.
KV:             DS      1       ;Epson for Vd.
```

```
Lsb8:           DS      1       ;Stator inductance (Lls+Lm)/4
Lsig:           DS      1       ;Combined inductance with sat. (appr. Lls+Llr)
Lsigr:          DS      1       ;Rated Lsig.
Rs:             DS      1       ;Stator resistance.
NUM:            DS      1       ;Auxiliary variable.
Asb2:           DS      1
Cs:             DS      1

; ----- Page 6, test variables
Iqave:          DS      1       ;Average Iqe
SIqH:           DS      1
SIqL:           DS      1
Weave:          DS      1       ;Average We
SWeH:           DS      1
SWeL:           DS      1
FLAG:           DS      1       ;Speed response to square wave comm.
Vs:             DS      1       ;Stator voltage magnitude.

; ---------- RULE BASE I - Kf COMPUTATION, E AND CE COMPUTATION
;                 USING P. 7!
                ORG     0
DVb4:           DS      1       ;Delta Vd/4.
DEMAX:          DS      1       ; = 32767 - 1024, Max. DE.
Wmax:           DS      1       ;Upper bound for We (32767 - Wb)
Imax:           DS      1       ;Upper bound for Iqe (32767 - Iqmax)
Wea:            DS      1       ;Wea is the auxiliar We (bounded)
Iqea:           DS      1       ;Iqea is the auxiliar Iqe (bounded)

RWb:            DS      1       ;1/Wb, scaled
MWH:            DS      1       ;Degree of membership HIGH for Wea.
MWL:            DS      1       ;Degree of membership LOW for Wea.
DMONE:          DS      1       ;Degree of membership one.

RIqr:           DS      1       ;1/Iqr, scaled.
MIH:            DS      1       ;Degree of membership HIGH for Iqea.
MIL:            DS      1       ;Degree of membership LOW for Iqea.

MRA:            DS      1       ;= MIN(MWH,MIH)
MRB:            DS      1       ;= MIN(MWL,MIH)
MRC:            DS      1       ;= MIN(MWH,MIL)
MRD:            DS      1       ;= MIN(MWL,MIL)

Kf1:            DS      1       ;Membership functions for Kf.
Kf2:            DS      1
Kf3:            DS      1

SUMPROD:        DS      1       ;Sum of product: MRi*Kfn
SUMMR:          DS      1       ;Sum of D.M.'s:MRA+..+MRD
Kf:             DS      1       ;Weight factor between Q and Vd errors.
DQE:            DS      1       ;Delta Q (error from Q model)
Err:            DS      1       ;Combined error.

e1min:          DS      1       ;-32676 + e1
EkcH:           DS      1       ;Clamped Error at time k
Ekc_1H:         DS      1
EkcL:           DS      1
Ekc_1L:         DS      1       ;  "     "   from time k-1
GC:             DS      1       ;Input gain for change in error.
DE:             DS      1       ;Original delta error.

A3:             DS      1       ;Coeficients for CE filters.
B3:             DS      1
```

```
C3:             DS      1
D3:             DS      1

xkc1:           DS      1       ;State variables for CE filter.
xkc2:           DS      1
xcl1:           DS      1       ;16 lsb of xkc1
xcl2:           DS      1       ;16 lsg of xkc2
CE:             DS      1       ;Filtered change in error, bounded by c3.
TEMP3:          DS      1       ;Auxiliary variable for p. 7.

; ---------- Rule base II variables and adjustable constants.
RE21:           DS      1       ; 1/(e2-e1), scaled.
RE32:           DS      1
RE3b2:          DS      1
RE4:            DS      1
RE54:           DS      1
RE65:           DS      1
RC1:            DS      1       ; 1/c1, scaled.
RC21:           DS      1                                               ;50
RC32:           DS      1
Ksmax:          DS      1       ;32767 - MaxKs (520) = 32247.
Ksmin:          DS      1       ;(32767 + MinKs(130))/2 = 16449
DKA:            DS      1       ;Rule A value.
DKB:            DS      1
DKC:            DS      1
DKD:            DS      1

ME1:            DS      1       ;Error membership value (left).
ME2:            DS      1
MC1:            DS      1       ;Change in error memb. value (left).
MC2:            DS      1
IJ:             DS      1       ;Offset for rule base look-up table.
IC:             DS      1       ;Row index.
JE:             DS      1       ;Column index.
DKs:            DS      1       ;Delta Ks.
TEMPi:          DS      1       ;Auxiliary variable.
GKs:            DS      1       ;Output gain.
KsH:            DS      1       ;High byte for slip gain.
KsL:            DS      1       ;Low byte  for slip gain.


; ---------------------------------------------------------------
;       RESET AND INTERRUPT VECTORS
; ---------------------------------------------------------------
;       ORG     0
;RESET: B       INIT    ;Reset vector:currently in RESMON
        ORG     4
INT1:   B       AINT    ;Channel A service interrupt routine
        ORG     6
INT2:   B       BINT    ;Channel B service interrupt routine
        ORG     24
INTT:   B       TINT    ;MAIN INTERRUPT (TIMER)
        ORG     26
RCV:    B       ININT   ;ADC service interrupt routine

; -----------------------------------------------------
;           INITIALIZE PROCESSOR
; -----------------------------------------------------
        ORG     1024            ;WITH RESMON ORG MUST BE 1024

INIT:   LDPK    0
        LALK    PRDC
```

```
        SACL    PRD             ;SET TIMER PERIOD

        SOVM
        SSXM                    ;set sign extension mode
        SPM     0               ;No shift in product register output
        FORT    0               ;Configure for 16 bit word serial data
        SXF                     ;Set output channel flag
        LACK    30              ;Enable RINT(16), AINT(2), BINT(4),TIMER(8)
        SACL    IMR             ;Set interrupt mask register
        LDPK    6
        LACK    1
        SACL    INCOMDATA
        OUT     INCOMDATA,INCOM         ;Clear M and T counters

; ---------- CONSTANT INITIALIZATION
        ZAC
        SACL    SSAMP
        LALK    DELWRCC
        SACL    DELWRC
        LALK    5
        SACL    SWT1
        LACK    50
        SACL    SWT0
        LALK    300
        SACL    SWT2
        ZAC
        SACL    FLAG

; ---------- Initialization of task handler and speed computation
        LDPK    6
        IN      MOLD,MCNT       ;Initialize MOLD=MCOUNT
        IN      TOLD,TCNT       ;   "     TOLD=TCOUNT
        ZAC
        SACL    DELWr0          ;Case of constant acceleration.
        SACL    Wr1
        SACL    Wr
        SACL    Wref
        SACL    WROLD
        SACL    OUTBFA
        SACL    OUTBFB
        SACL    STP
        LALK    SLIPGAIN
        SACL    Kslip
        LALK    28377
        SACL    RT34

; ---------- BUFFER INITIALIZATION
        LACK    3
        SACL    BUFAH
        LALK    0FE00H
        SACL    BUFAL
        ZAC
        SACL    VAR1            ;Selection of variable to be stored
        SACL    VAR2
        SACL    ZERO
        LACK    1
        SACL    ONE

; ---------- FILTER 1 INIT.: SLIP GAIN COMPUTATION UNDER VAR. FLUX
        LALK    A1C             ;Initialize coeficients
        SACL    A1
        LALK    B1C
```

```
        SACL    BI
        LALK    C1C
        SACL    CI
        LALK    DIC
        SACL    DI

        LALK    A2C             ;Initialize coeficients for Wr filter
        SACL    A2
        LALK    B2C
        SACL    B2
        LALK    C2C
        SACL    C2
        LALK    D2C
        SACL    D2
        LALK    Iderated        ;Initialize Ide (always to rated value)
        SACL    Ide

; ---------- INITIALIZE VARIABLES FOR PI CONTROLLER AND SPEED FILTER
        ZAC
        SACL    WRI             ;Original feedback speed
        SACL    WR              ;Filtered feedback speed
        SACL    XWK
        SACL    WREF
        SACL    Iqe
        SACL    IQH
        SACL    IQL
        SACL    EKI
        LALK    K1C
        SACL    KI
        LALK    K2C
        SACL    K2
        LALK    IQMV
        SACL    IQMAX           ;Store iqmv into IQMAX
        LALK    32767
        SUB     IQMAX           ;
        SACL    ILIM            ;ILIM = 32767 - IQMAX

; ---------- INITIALIZE VARIABLES AND CONSTANTS FOR HOST/DSP COMM.
        LALK    29320
        SACL    KWREF           ;rpm to binary coef.
        LALK    4578
        SACL    KBR             ;Binary to rpm coef.

; ---------- INITIALIZE VARS AND CONSTANTS FOR SLIP GAIN PROJECT
; ---------- DELTAC ROUTINE (p. 6)
        LALK    3224            ;Coefficients for temp. compensation
        SACL    AV              ; in Vds and Vqs sensing.
        LALK    -465
        SACL    BVs
        LALK    18919
        SACL    RTI23           ;RTI23 = 1 / SQRT(3).
        LALK    17731
        SACL    Lsb8
        LALK    13548
        SACL    Lsigr           ;Rated Lsig.
        SACL    Lsig            ;Actual Lsig.
        LALK    725
        SACL    Rs
        LACK    4
        SACL    KQ              ;KQ = 0.5 VA or 4 bits.
        LALK    131             ;KV = 1.6 V or 262 bits.
        SACL    KV              ;KV = 0.1 V or 16 bits.
```

```
        LALK    24363           ;Coefficients for Lsig saturation effect.
        SACL    Asb2
        LALK    -31147
        SACL    Bsb2
        LALK    20096
        SACL    Cs

; ---------- FILTER ROUTINE (Vds AND Vqs FILTERS)
        ZAC
        SACL    FSG             ;Flag for slip gain process.
        SACL    xkd1
        SACL    xkd2
        SACL    xkq
        SACL    xd1l
        SACL    xd2l
        SACL    xql
        LALK    12000           ;Now using isolation amp. circuit.
        SACL    Talv            ;Measured time delay = 72.2 Micro sec.
        LALK    10000           ;Stator voltage magnitude (Vs(0)).
        SACL    Vs

; ---------- VARIABLES AND CONSTANTS FOR RULE BASE I
        LDPK    7               ;Uses page 7
        LALK    17019
        SACL    Wmax            ;Wmax = 32767 - Wb(=15748).
        LALK    7720
        SACL    Imax            ;Imax = 32767 - Iqrated (=25047).
        LALK    3270
        SACL    DMONE           ;DMONE = degree of membership 1.
        LALK    6804
        SACL    RWb             ;RWb = 1/Wb, scaled.
        LALK    4278
        SACL    RIqr            ;RIqr = 1/Iqrated.
        LALK    26214
        SACL    Kf1             ;Kf1 = 0.8 (26214)
        LALK    29490
        SACL    Kf2             ;Kf2 = 0.9 (29490)
        LALK    32112
        SACL    Kf3             ;Kf3 = 0.98 (32112)
        LALK    -13107
        SACL    elmin           ;elmin = -32768 - elmin (= -0.6*32768)
        LALK    18431           ; For GC = 400, 18431.
        SACL    GC              ;GC = remainder of 400/256.
        LALK    31743           ;32686          ;
        SACL    DEMAX           ;DEMAX = 32767 - max DE ( = 1024).

        LALK    A3C             ;Initialize coeficients for CE filter
        SACL    A3
        LALK    B3C
        SACL    B3
        LALK    C3C
        SACL    C3
        LALK    D3C
        SACL    D3

        ZAC
        SACL    xkc1            ;State var. for CE filter.
        SACL    xkc2
        SACL    xc1l
        SACL    xc12            ;16 slb of xkc1, xkc2.

; ---------- VARIABLES AND CONSTANTS FOR RULE BASE II.
```

```
        LALK    10900           ;REmn = I / (em - en)
        SACL    RE21
        LALK    13625
        SACL    RE32
        LALK    27250
        SACL    RE3b2
        LALK    32700
        SACL    RE4
        LALK    8175
        SACL    RE54
        LALK    6540
        SACL    RE65

        LALK    16350           ;RCmn = I / (cm - cn).
        SACL    RC1
        LALK    8175
        SACL    RC21
        LALK    8175
        SACL    RC32

        LALK    32247
        SACL    Ksmax           ;Ksmax = 32767 - maxKs.
        LALK    16449
        SACL    Ksmin           ;Ksmin = (32767+Ksmin)/2.

        LALK    12452           ;Original gain = 82.
        SACL    GKs             ;Output gain.

        LALK    SLIPGAIN        ;Initialize Ksh with rated slip gain.
        SACL    KsH
        ZAC
        SACL    KsL

; ---------------------------------------------------------------------------
;  BACKGROUND LOOP - HOST COMMUNICATION AND IDLE TIME
; ---------------------------------------------------------------------------
MAIN:   EQU     $
        EINT
        LDPK    6               ;Write selected variable to HOST
        OUT     DISP,WRITEHOST  ;VAR ---> WRITEHOST (PA15)

; ---------------------------------------------------------------------------
;           HOST/DSP16 COMMUNICATION ROUTINE
; ---------------------------------------------------------------------------
NCLUD HOSTDSP.ASM

; ---------------------------------------------------------------
;                 MAIN INTERRUPT ROUTINE
; ---------------------------------------------------------------
;
;---------- CONTEXT SAVE
;                                   AR7 IS THE STACK POINTER ( AR7=124!)

TINT:   LARP    AR7
        LARK    AR7,124
        MAR     *-
        SST1    *-
        SST     *-
        EINT
        SACH    *-
        SACL    *-
        SPH     *-
```

```
        SPL     *-
        MPYK    1
        SPL     *-
        LDPK    6

;-------------------------------------------------------------------------------
;  PERFORM SLIP GAIN TUNING ROUTINES WITH Ts = 0.2 ms.
;-------------------------------------------------------------------------------
        CALL    IVR             ;Reverse vector rotation (Vde, Vqe comp.)
        CALL    FILTER          ;Vde and Vqe filters.

;-------------------------------------------------------------------------------
;          PERFORM FEEDBACK SPEED COMPUTATION
;-------------------------------------------------------------------------------
        CALL    SPEED
        LAC     Wr1
        SACL    Wr

;-------------------------------------------------
;          PI SPEED CONTROL
;-------------------------------------------------
TS1MS:  LDPK    6               ;1 ms sampling time routines
        LAC     SWT1            ;Check for the sampling time
        SUBK    1
        SACL    SWT1            ;Decrement counter SWT1
        BGZ     TS1RET
        LACK    5               ;5<-->1ms
        SACL    SWT1            ;Reload counter if zero
        CALL    PI              ;PI speed controller

;-------------------------------------------------------------------------------
;  PERFORM SLIP GAIN TUNING ROUTINES WITH Ts = 1 ms.
;-------------------------------------------------------------------------------
        LAC     FSG
        BZ      TS1RET          ;Slip gain tuning is active if FSG = 1.
        CALL    DELTAC          ;Delta Q and delta Vd computation.
        CALL    RBI             ;KF, E and CE computation.
        CALL    RBII            ;DKs, Ks computation.
TS1RET: EQU     $

;-------------------------------------------------------------------------
;       PERFORM BASIC VECTOR CONTROL FUNCTIONS
;-------------------------------------------------------------------------
        CALL    VECROT ;Original program: slip calc. + v.r

;-------------------------------------------------
;      OUTPUT VARIABLES (FOR DAC)
;-------------------------------------------------
        SPM     1
        LT      Ias             ;Compensates for DAC gain mismatches
        MPYK    203
        PAC
        ADDH    Ias
        SACH    Ias
        LAC     Ias
        ADDK    64              ;Offset compensation for DAC A
        SACL    OUTBFA          ;Channel A of DAC

        LT      Ibs             ;Compensates for DAC gain mismatches
        MPYK    139
        PAC
        ADDH    Ibs
```

```
        SACH    Ibs
        LAC     Ibs
        ADDK    115             ;Offset compensation for DAC B
        SACL    OUTBFB          ;Channel B of DAC
        CALL    AVE16           ;1.024 sec average for Vdef, Vqef  (monitoring ).

; ---------------------------------------------------
;               BUFFER STORAGE
; ---------------------------------------------------
; ---------- SELECT THE VARIABLE TO BE STORED IN CHANNEL A
        SPM     1                               ;****************************
        LAC     VAR1                            ;*VAR1 = 1  STORE:  Ekch    *
        LDPK    6                               ;*                          *
        SUBK    1               ;Case VAR1 =  1 ;*         2         Iqe    *
        BNZ     VA2                             ;*         3         Wref   *
        LDPK    7                               ;****************************
        LAC     Ekch
        NEG
        B       VAEND

VA2:    SUBK    1               ;Case VAR1 = 2
        BNZ     VA3
        LAC     Iqe
        NEG
        B       VAEND

VA3:    SUBK    1               ;Case VAR1 = 3
        BNZ     VBSTART         ;Wrong value for VAR1; check VAR2
        Wref

VAEND:  LDPK    6
        SACL    DISP
        SACL    BFADATA         ;Store selected variable into CH. A

; ---------- SELECT THE VARIABLE TO BE STORED IN CHANNEL B
;
VBSTART:LDPK    6                       ;****************************
        LAC     VAR2                    ;*VAR2 = 1  STORE:  Kslip   *
        SUBK    1       ;Case VAR2 =  1 ;*         2         CE     *
        BNZ     VB2                     ;*         3         Iqe    *
        LAC     Kslip,5                 ;****************************
        NEG
        B       VBEND

VB2:    SUBK    1               ;Case VAR2 = 2
        BNZ     VB3
        LDPK    7
        LAC     CE
        B       VBEND

VB3:    SUBK    1               ;Case VAR2 = 3
        BNZ     VAREND          ;Value of VAR2 is not valid; ignore it.
        LDPK    6
        LAC     Iqe

VBEND:  LDPK    6
        SACL    BFBDATA
        SACL    DISP

; ---------------------------------------------------------------
;       BUFFER STORAGE SUBROUTINE
; ---------------------------------------------------------------
```

```
VAREND:         LDPK    6
; ---------------------------------------------------------------
        LAC     SWT0            ;Check for the sampling time of Buffer Storage.
        SUBK    1
        SACL    SWT0            ;Decrement counter SWT0
        BGZ     BUFRET          ;By-pass if not required
        LACK    50              ;SWT0 = 10 ms (variable) 50 x 0.2ms = 10ms
        SACL    SWT0            ;Reload counter if zero and
        CALL    BUFSTR          ;Perform buffer storage
BUFRET: EQU     $

;--------- CONTEXT RESTORE
        LARP    AR7
        MAR     *+
        MAR     *+
        LT      *-
        MPYK    1
        LT      *+
        MAR     *+
        LPH     *+
        ZALS    *+
        ADDH    *+

        DINT
        LDPK    0
        SST1    SSTEMP
        BIT     SSTEMP,0bh
        BBNZ    PPPP
        LDPK    6
        LST     *+
        EINT
        RET
PPPP:   LDPK    6
        LST     *+
        EINT
        RET

; ---------------------------------------------------------------
;  LIST OF INCLUDED SUBROUTINES
; ---------------------------------------------------------------
NCLUD ADCSG.ASM                 ;ADC INTERRUPT SERVICE ROUTINE
NCLUD BUFSTO.ASM                ; BUFFER STORAGE SUBROUTINE
NCLUD DAC.ASM                   ;OUTPUT INTERRUPT SERVICE ROUTINE
NCLUD PICOMP.ASM                ;PI SPEED CONTROLLER
NCLUD SPEED.ASM                 ; SPEED FEEDBACK SUBROUTINE
NCLUD FOFW.ASM                  ;First order filter for speed, FOFW
NCLUD SIN.ASM                   ;TABLE FOR SIN/COS COMPUTATION
NCLUD VR.ASM                    ; BASIC VECTOR CONTROL FUNCTIONS

; ---------------------------------------------------------------
;         SLIP GAIN TUNING FUNCTIONS
; ---------------------------------------------------------------
NCLUD IVR.ASM                   ;Reverse vector rotation.
NCLUD FILTER.ASM                ;Vde and Vqe filters.
NCLUD DELTAC.ASM                ;Delta Q and delta Vd computation.
NCLUD RBI.ASM                   ;Kf, E and CE computation.
NCLUD RBII.ASM                  ;DKs, Ks computation.
NCLUD AVE16.ASM                 ;16 ms average for Vdef and Vqef.
NCLUD VSMOD.ASM                 ;Stator voltage amplitude computation.
; ------------------------- END OF PROGRAM -------------------------
```

```
; ------------------------------------------------------------------
;       Computation of DELTA Q and DELTA Vd
; ------------------------------------------------------------------
;
DELTAC: SPM     1
        LDPK    6
        SOVM

; ---------- Compute QE, actual Q from measurements
        LT      Vqef
        MPY     Ide
        LTP     Vdef            ;ACC = Vqef*Ide; TR = Vdef
        MPY     Iqe
        SPAC
        SACH    QE

; ---------- Compute QEC, command Q from reference model
        SQRA    Ide
        LTP     Lsb8            ;TR = Ls/8, ACCH = Ide^2
        SACH    TEMP0           ;TEMP0 = Ide^2
        MPY     TEMP0
        PAC
        SACH    TEMP0,3         ;TEMP0 = 8* Ls/8 *Ide^2
        SQRA    Iqe
        LTP     Lsig            ;TR = Lsig, ACC = Iqe^2
        SACH    TEMP1           ;TEMP1 = Iqe^2
        MPY     TEMP1
        PAC
        ADDH    TEMP0           ;ACCH = Ls*Ide^2+Lsig*Iqe^2
        SACH    TEMP1
        LT      We
        MPY     TEMP1
        PAC
        SACH    QEC             ;QEC =  We*(Ls*Ide^2+Lsig*Iqe^2)

        ADDH    KQ              ;Qb = QEC + KQ ;KQ = constant
        SACH    Qb

; ---------- Compute DELTA Q, by division
        LAC     QEC
        SUB     QE
        SACL    NUM             ;NUM = QEC-QE

; ---------- Clamp |QEC - QE| to Qb
        LALK    32767
        SUB     Qb
        SACL    TEMP1
        ZALH    NUM
        ADDH    TEMP1
        SUBH    TEMP1
        SUBH    TEMP1
        ADDH    TEMP1

; ---------- Divide NUM BY Qb to get DQE (Fractional division)
;       Here, the sign of the result is given by NUM only,
;           since Qb is always positive
DST1:   ZALH    NUM
        ABS                     ;Make numerator positive
        RPTK    14
        SUBC    Qb              ;Result is in low ACC
        SACL    TEMP1
        LAC     NUM
```

```
        BGEZ    DONE1           ;Done if sign is positive
        LAC     TEMP1
        NEG
        SACL    TEMP1           ;Negate quotient if NUM is negative

DONE1:  LAC     TEMP1
        LDPK    7
        SACL    DQE             ;DQE is now in p. 7!

        LDPK    6

; ---------- Introduce saturation effect on Lsig.
;             xsr = A Is^4 + B Is^2 + C.
;         Lsig = Lsigr + xsr*Lsigr
        SQRA    Ide
        ZAC                     ;PR = Ide^2.
        SQRA    Iqe             ;ACCH = Ide^2, PR = Iqe^2.
        APAC
        SACH    TEMP0           ;Is2 = Ide^2 + Iqe^2 = TEMP0.
        SQRA    TEMP0
        PAC
        SACH    TEMP1           ;Is4 = Is^4 = TEMP1.

        ZALH    Cs
        MPY     Bsb2            ;TR = Is2 has not changed.
        APAC                    ;PR = (Bs/2) Is^2.
        LTA     TEMP1           ;ACCH = B*Is^2 + C.
        MPY     Asb2            ;TR = Is4.
        APAC                    ;PR = (As/2) Is^4.
        LTA     Lsigr           ;ACCH = A Is^4 + B Is^2 + C = xsr = TEMP2.
        SACH    TEMP2
        ZALH    Lsigr           ;TR = Lsigr.
        MPY     TEMP2
        APAC                    ;ACCH = Lsigr + xsr*Lsigr = Lsig.
        SACH    Lsig            ;Saturation compensated Lsig.

; ---------- Perform computation of Vdec, commanded Vd
        LT      We
        MPY     Lsig            ;TR = We.
        LTP     Iqe             ;TR = Iqe, ACC = We*Lsig
        SACH    TEMP0           ;TEMP0 = We*Lsig
        MPY     TEMP0
        LTP     Ide             ;ACCH = We*Lsig*Iqe = TEMP0; TR = Ide
        SACH    TEMP0
        MPY     Rs
        PAC                     ;ACCH = Rs*Ide
        SUBH    TEMP0
        SACH    Vdec            ;Vdec = Rs*Ide - TEMP0
        LAC     TEMP0
        ABS
        ADD     KV              ;KV is a constant
        SACL    Vb,2            ;Vb = 4*(KV + ABS(We*Lsig*Iqe))

; ---------- Compute DELTA Vd, by division
        LAC     Vdef
        SUB     Vdec            ;Sign reversal is intentional!
        SACL    NUM             ;NUM = Vdef - Vdec
        LALK    32767
        SUB     Vb              ;Clamp |NUM| to 4*Vb
        SACL    TEMP1           ; to ensure proper fractional division.
        ZALH    NUM
        ADDH    TEMP1
```

```
          SUBH    TEMP1
          SUBH    TEMP1
          ADDH    TEMP1
          SACH    NUM

; ---------- Divide NUM BY Vb(x4) to get DVb4 (Fractional division)
;          Here, the sign of the result is given by NUM only,
;               since Vb is always positive
          ZALH    NUM
          ABS                     ;Make numerator positive
          RPTK    14
          SUBC    Vb              ;Result is in low ACC
          SACL    TEMP1
          LAC     NUM
          BGEZ    DONE2           ;Done if sign is positive
          LAC     TEMP1
          NEG
          SACL    TEMP1           ;Negate quotient if NUM is negative

DONE2:    LAC     TEMP1
          LDPK    7
          SACL    DVb4
          LDPK    6
          RET


; ---------------------------------------------------------------
;        FILTER FOR Vde AND Vqe
; ---------------------------------------------------------------
;          fcf = 53.05 Hz (TAL = 3 ms) ,  fsamp = 5 KHz
;          y(k) = c x(k) + d u(k)
;          x(k+1) = a x(k) + b u(k)
;
;          . Two cascade first order filters are used for Vde
;          . One filter is used for Vqe
;          . Discrete version of ECALC (first part of)
; ---------------------------------------------------------------
FILTER:   SPM     1
          LT      xkd1
          MPY     C1
          LTP     Vde             ;ACC = c x(k) , T = u(k)
          MPY     D1
          APAC                    ;ACC = c x(k) + d u(k)
          SACH    Vde1            ;Vde after first filter
          MPY     B1              ;T = u(k) has not changed
          LTP     xkd1            ;ACC = b u(k) , T = x(k)
          MPY     A1
          APAC                    ;ACC = a x(k) + b u(k)
          ADDS    xd11            ;Add 16 lsb from last iteration
          SACH    xkd1            ;Actually x(k+1) ---> x(k) for next iteration
          ANDK    65280           ;Get upper half of ACCL.
          SACL    xd11            ;Save 16 lsb for next iteration.

; ---------- Get filtered version of Vde
          LT      xkd2
          MPY     C1
          LTP     Vde1            ;ACC = c x(k) , T = u(k)
          MPY     D1
          APAC                    ;ACC = c x(k) + d u(k)
          SACH    Vdef
          MPY     B1              ;T = u(k) has not changed
          LTP     xkd2            ;ACC = b u(k) , T = x(k)
```

```
          MPY     A1
          APAC                    ;ACC = a x(k) + b u(k)
          ADDS    xd2l            ;Add 16 lsb from last iteration
          SACH    xkd2            ;Actually x(k+1) ---> x(k) for next iteration
          ANDK    65280           ;Get upper half of ACCL.
          SACL    xd2l            ;Save 16 lsb for next iteration.

; ---------- Get filtered Vqe (1st order only)
          LT      xkq
          MPY     C1
          LTP     Vqe             ;ACC = c x(k) , T = u(k)
          MPY     D1
          APAC                    ;ACC = c x(k) + d u(k)
          SACH    Vqef            ;Filtered Vqe
          MPY     B1              ;T = u(k) has not changed
          LTP     xkq             ;ACC = b u(k) , T = x(k)
          MPY     A1
          APAC                    ;ACC = a x(k) + b u(k)
          ADDS    xql             ;Add 16 lsb from last iteration
          SACH    xkq             ;Actually x(k+1) ---> x(k) for next iteration
          ANDK    65280           ;Get upper half of ACCL.
          SACL    xql             ;Save 16 lsb for next iteration.
          RET



; ------------------------------------------------------------------------------------------
; REVERSE VECTOR ROTATOR FOR Vds AND Vqs COMPUTATION
; ------------------------------------------------------------------------------------------
; ---------- Correct phase-shift due to analog filters
IVR:      SPM     0
; ----------------------------------------------------
          LDPK    6
          ROVM
          ZALH    THETAH          ;Angle from forward vector rotator
          ADDS  THETAL
          LT      We
          MPY     Talv            ;Correcting angle is proportional
          SPAC                    ; to We, when phase shift is small.
          SACH    TEMP0           ;Current Taleq = 9145 bits (457.3 Ms)

; ---------- Compute shift-compensated unit vectors
          SSXM
          LT      TEMP0           ;TEMP0 = Corrected angle
          MPYK    2048
          PAC
          SACH    TEMP0           ;Scaling to access SIN table
          LAC     TEMP0
          BGEZ    POSIND
          ADLK    1024            ;If negative, add 180 degrees
          ANDK    1023            ;Extract the first seven bits only.

NESIND:   SOVM
          ADLK    SINTBL          ;Add address offset
          TBLR    SIND            ;Retrieve sin(thetac)
          ADLK    TBL90
          TBLR    COSD            ;Retrieve cos(thetac)
          LAC     SIND
          NEG
          SACL    SIND            ;Reverse sign of vectors
          LAC     COSD
```

```
          NEG
          SACL    COSD
          B       SINED

POSIND:   SOVM                    ;Case of positive angle
          ADLK    SINTBL
          TBLR    SIND
          ADLK    TBL90
          TBLR     COSD

SINED:    EQU     $

; ---------- Make ADC offset corrections and temp. compensation
;         for Vds and Vqs sensing.
          SPM     1               ;From this point on, use PM = 1.
          LAC     Vqs
          ADLK    164
          SACL    TEMP1           ;Offset compensation.
          LAC     Vds
          SUBK    40
          SACL    TEMP2

          LT      AV              ;Temperature effect compensation.
          MPY     We
          PAC
          ADDH    BVs
          SACH    TEMP0           ;TEMP0 = Coefficient of correction for
                                  ;       temp. effect on Vds and Vqs.
          LT      TEMP0           ;TR = temp. coeff.
          MPY     TEMP1
          PAC
          ADDH    TEMP1
          SACH    TEMP1           ;TEMP1 = Vqs*(1 + coeff)
          MPY     TEMP2           ;TR = temp. coeff.
          PAC
          ADDH    TEMP2
          SACH    TEMP2           ;TEMP2 = Vds*(1 + coeff)

; ---------- Stationary to synchronous frame convertion
;         Vqe= Vqs*COS-Vds*SIN
;         Vde= Vqs*SIN+Vds*COS
          LT      COSD
          MPY     TEMP1
          LTP     SIND            ;ACC = Vqs*COSD
          MPY     TEMP2
          SPAC                    ;ACC = Vqs*COSD - Vds*SIND = Vqe
          SACH    Vqe,1           ;Uses a new scale for voltage.!
          MPY     TEMP1           ;TR = SIND
          LTP     COSD            ;ACC = Vqs*SIND; TR = COSD
          MPY     TEMP2
          APAC                    ;ACC = Vqs*SIND + Vds*COSD
          SACH    Vde,1           ;Using new scale for voltage.
          RET


; ------------------------------------------------------------------------------
;       RULE BASE I - Kf COMPUTATION
; ALSO E (Error) AND CE (Change in Error) COMPUTATION
; ------------------------------------------------------------------------------
;     NOTE: VARIABLES ARE DEFINED IN  P. 7
RBI:      SPM     1
          LDPK    6
```

```
        LAR     AR1,We          ;Copy We and Iqe from p. 6 into p. 7.
        LAR     AR2,Iqe
        LDPK    7
        SAR     AR1,Wea
        SAR     AR2,Iqea

        SOVM                    ;Use overflow mode to clamp We, Iqe
        ZALH    Wea
        ABS                     ;Only module of We is relevant.
        ADDH    Wmax            ;Clamp We to Wb = 2200 rpm.
        SUBH    Wmax
        SACH    Wea             ;Wea is the auxiliar We
        ZALH    Iqea
        ABS                     ;Only module of Iqe is relevant.
        ADDH    Imax            ;Clamp Iqe to Iqrated = 19.2 A.
        SUBH    Imax
        SACH    Iqea            ;Iqea is the auxiliar Iqe

; ---------- Get degree of membership for Wea and Iqea
        LT      Wea
        MPY     RWb             ;RWb is 1/Wb, scaled.
        PAC                     ;ACC = MWH, D.M. HIGH for Wea.
        SACH    MWH
        LAC     DMONE
        SUB     MWH             ;ACC = 1 - MWH = MWL
        SACL    MWL

        LT      Iqea
        MPY     RIqr            ;RIqr is 1/Iqr, scaled.
        PAC                     ;ACC = MIH, D.M. of right MF for Iqea
        SACH    MIH
        LAC     DMONE
        SUB     MIH             ;ACC = 1 - MIH = MIL
        SACL    MIL

; ---------- Perform rule base evaluation (Min operator)
        LAC     MWH             ;MRA = MIN(MWH,MIH)
        SACL    MRA             ;Assume MRA = MWH
        SUB     MIH             ;ACC = MP1-MIH
        BLEZ    NXT1            ;Assumption is correct if ACC<=0
        LAC     MIH             ;No: MP1>MIH. Make MRA=MIH
        SACL    MRA

NXT1:   LAC     MWL             ;MRB = MIN(MWL,MIH)
        SACL    MRB             ;Assume MRB = MWL
        SUB     MIH             ;ACC = MWL-MIH
        BLEZ    NXT2            ;Assumption is correct if ACC<=0
        LAC     MIH             ;No: MWL>MIH. Make MRB=MIH
        SACL    MRB

NXT2:   LAC     MWH             ;MRC = MIN(MWH,MIL)
        SACL    MRC             ;Assume MRC = MWH
        SUB     MIL             ;ACC = MWH-MIL
        BLEZ    NXT3            ;Assumption is correct if ACC<=0
        LAC     MIL             ;No: MWH>MIL. Make MRC=MIL
        SACL    MRC

NXT3:   LAC     MWL             ;MRD = MIN(MWL,MIL)
        SACL    MRD             ;Assume MRD = MWL
        SUB     MIL             ;ACC = MWL-MIL
        BLEZ    NXT4            ;Assumption is correct if ACC<=0
        LAC     MIL             ;No: MWL>MIL. Make MRD=MIL
```

```
        SACL    MRD
NXT4:   EQU     $

; ------------------------------------------------------
; ---------- DEFUZZIFICATION BY HEIGHT METHOD
; ---------- Get sum of truth value of fired rules (MRA+MRB+MRC+MRD)
        LAC     MRA
        ADD     MRB
        ADD     MRC
        ADD     MRD
        SACL    SUMMR           ;SUMMR = MRA+MRB+MRC+MRD

;---------- Get sum of products: MRi*Kfn, i=A,B,C,D, n=1,2,3
        LT      MRA
        MPY     Kf2
        LTP     MRB             ;ACC = MRA*Kf2, TR = MRB
        MPY     Kf3
        LTA     MRC
        MPY     Kf1
        LTA     MRD
        MPY     Kf2
        APAC
        SACH    SUMPROD         ;SUMPROD=MRA*Kf2+....+MRD*Kf2

; ---------- Divide SUMPROD BY SUMMR do get Kf (Fractional division)
;       Here, the sign of the result is defined by SUMPROD only
;       since SUMMR  is by def. always positive
        ZALH    SUMPROD
        ABS                     ;Make numerator positive
        RPTK    14
        SUBC    SUMMR           ;Result is in low ACC
        SACL    Kf
        LAC     SUMPROD
        BGEZ    DIVDONE         ;Done if sign is positive
        ZAC
        SUB     Kf
        SACL    Kf              ;Negate quotient if negative
DIVDONE:EQU     $

; ---------- Compute Error (E) and Change in Error (CE).
        SPM     0               ;PM = 0 introduces GE = 0.5 automatically!
        LALK    -32767
        ADD     Kf
        SACL    TEMP3,2         ;TEMP3 = -4*(1 - Kf). As (1 - Kf)<0.2
                                ; no overflow occurs.
        LT      Kf
        MPY     DQE
        ZAC
        LTS     TEMP3           ;ACC = -0.5*Kf*DQE, TR = -4*(1 - Kf).
        MPY     DVb4            ;DVb4 = DVde / 4.
        APAC
        SACH    Err             ;Err = -0.5*(Kf*DQ + 4*(1-Kf)*DVb4)
        SPM     1               ;Restores PM to its normal value.

; ---------- Now impose lower (e1) bound to Err
;       (Upper bound e6 has been enforced by OVM = 1)
        SUBH    e1min           ;e1min = -32767 + e1.
        ADDH    e1min
        SACH    EkcH            ;Ekc = Err, if Err >  e1.
        SACL    EkcL            ;Stores in 32 bit to improve precision.

; ---------- Compute Change in Error
```

```
        SUBS    Ekc_1L
        SUBH    Ekc_1H          ;Get delta error in extended precision.
        ADDH    DEMAX           ;Use saturation features to clamp DE to 1024
        SUBH    DEMAX           ;DEMAX = 32767 - 1024 = 32686
        SUBH    DEMAX
        ADDH    DEMAX
        SACH    DE,5            ;Define DE = (Ekc - Ekc_1)*2^5.
        DMOV    EkcH            ;Ekc --> Ekc_1.
        DMOV    EkcL            ;Ekc --> Ekc_1.

; ---------- Pass DE through filters to get CE (final change in error)
        LT      xkc1
        MPY     C3
        LTP     DE              ;ACC = c x(k) , T = u(k)
        MPY     D3
        APAC                    ;ACC = c x(k) + d u(k)
        SACH    TEMP3           ;DE after first filter
        MPY     B3              ;T = u(k) has not changed
        LTP     xkc1            ;ACC = b u(k) , T = x(k)
        MPY     A3
        APAC                    ;ACC = a x(k) + b u(k)
        ADDS    xc1l            ;Add 16 lsb from last iteration
        SACH    xkc1            ;Actually x(k+1) ---> x(k) for next iteration
        ANDK    65280           ;Get upper half of ACCL.
        SACL    xc1l            ;Save 16 lsb for next iteration.

; ---------- Second  filter for DE
        LT      xkc2
        MPY     C3
        LTP     TEMP3           ;ACC = c x(k) , T = u(k)
        MPY     D3
        APAC                    ;ACC = c x(k) + d u(k) = filtered DE
        ABS                     ;Only module of DE is relevant.
        SACH    CE,5            ;CE is the filtered version of DE * 2^5.
                                ;CE = 1024 * (E(k) - E(k-1))
        MPY     B3              ;T = u(k) has not changed
        LTP     xkc2            ;ACC = b u(k) , T = x(k)
        MPY     A3
        APAC                    ;ACC = a x(k) + b u(k)
        ADDS    xc12            ;Add 16 lsb from last iteration
        SACH    xkc2            ;Actually x(k+1) ---> x(k) for next iteration
        ANDK    65280           ;Get upper half of ACCL.
        SACL    xc12            ;Save 16 lsb for next iteration.
        RET


; ------------------------------------------------------------------------------
;           RULE BASE II - DELTA Ks COMPUTATION
; ------------------------------------------------------------------------------
RBII:   SOVM                    ;Use overflow mode troughout RBII
        SPM     1
        LDPK    7               ;Uses page 7 throughout.

; ---------- Compute Degree of Membership for ERROR (E)
        LAC     EkcH
        BGEZ    EGEZ
        SBLK    e3              ;-------- Case E<0
        SACL    TEMPi,1         ;Check if it is in the interval (e3,0).
        BLZ     NE1             ;Here TEMPi = 2 * (E - e3) to compensate for RE3b2.
        LT      RE3b2           ;Case e3<E<0
        LACK    3               ;Interval index J = 3
        SACL    JE
```

```
        B       Edone

NE1:    LAC     EkcH            ;Compare E with e2
        SBLK    e2
        BLZ     NE2
        LT      RE32            ;Case e2 < E < e3
        SACL    TEMPi
        LACK    2               ;Interval index J = 2.
        SACL    JE
        B       Edone

NE2:    LAC     EkcH
        SBLK    e1
        LT      RE21            ;Case e1 < E < e2
        SACL    TEMPi
        LACK    1               ;Interval index J = 1
        SACL    JE
        B       Edone


EGEZ:   SBLK    e5              ;------ Case E > 0
        BLZ     POS1            ;Test if e5 < E < e6
        SACL    TEMPi           ;Case e5 < E < e6
        LACK    6
        SACL    JE              ;Interval index J = 6
        LT      RE65
        B       Edone

POS1:   LAC     EkcH
        SBLK    e4
        BLZ     POS2            ;Test if e4 < E < e5
        SACL    TEMPi           ;Yes
        LACK    5
        SACL    JE              ;Interval index J = 5
        LT      RE54
        B       Edone

POS2:   LACK    4               ;Case  0 < E <e4
        SACL    JE              ;Interval index J = 4
        LT      RE4
        LAC     EkcH
        SACL    TEMPi

Edone:  MPY   TEMPi
        PAC                     ;ACC = ME2, D.M. of right MF for E
        SACH    ME2
        LAC     DMONE
        SUB     ME2             ;ACC = 1 - ME2 = ME1
        SACL    ME1

; ---------- Compute DM for Change in Error (CE).
        LAC     CE
        SBLK    ce2             ;CE > 0 by definition.
        BLZ     PCE1            ;Test if ce2 < CE < ce3
        SACL    TEMPi           ;Case e5 < E < e6
        LACK    1
        SACL    IC              ;Interval index I = 1.
        LT      RC32
        B       CEdone

PCE1:   LAC     CE
        SBLK    ce1
```

```
        BLZ     PCE2            ;Test if ce1 < CE < ce2
        SACL    TEMPi           ;Yes
        LACK    2
        SACL    IC              ;Interval index I = 2
        LT      RC21
        B       CEdone

PCE2:   LACK    3               ;Case  0 < CE < ce1
        SACL    IC              ;Interval index I = 3
        LT      RC1
        LAC     CE
        SACL    TEMPi

CEdone: MPY     TEMPi
        PAC                     ;ACC = MC2, D.M. of right MF for CE.
        SACH    MC2
        LAC     DMONE
        SUB     MC2             ;ACC = 1 - MC2 = MC1
        SACL    MC1

; ---------- Perform rule base evaluation (Min operator)
        LAC     ME1             ;MRA = MIN(ME1,MC2)
        SACL    MRA             ;Assume MRA = ME1
        SUB     MC2             ;ACC = ME1-MC2
        BLEZ    NXK1            ;Assumption is correct if ACC<=0
        LAC     MC2             ;No: ME1>MC2. Make MRA=MC2
        SACL    MRA

NXK1:   LAC     ME2             ;MRB = MIN(ME2,MC2)
        SACL    MRB             ;Assume MRB = ME2
        SUB     MC2             ;ACC = ME2-MC2
        BLEZ    NXK2            ;Assumption is correct if ACC<=0
        LAC     MC2             ;No: ME2>MC2. Make MRB=MC2
        SACL    MRB

NXK2:   LAC     ME1             ;MRC = MIN(ME1,MC1)
        SACL    MRC             ;Assume MRC = ME1
        SUB     MC1             ;ACC = ME1-MC1
        BLEZ    NXK3            ;Assumption is correct if ACC<=0
        LAC     MC1             ;No: ME1>MC1. Make MRC=MC1
        SACL    MRC

NXK3:   LAC     ME2             ;MRD = MCN(ME2,MC1)
        SACL    MRD             ;Assume MRD = ME2
        SUB     MC1             ;ACC = ME2-MC1
        BLEZ    NXK4            ;Assumption is correct if ACC<=0
        LAC     MC1             ;No: ME2>MC1. Make MRD=MC1
        SACL    MRD
NXK4:   EQU     $

; ---------- Get control signal for each of the 4 fired rules
; ---------- Compute offset for look-up table. (IJ)
        LAC     IC,3            ;ACC = 8*IC
        SUB     IC              ;ACC = 7*IC
        ADD     JE
        SUBK    8               ;ACC = 7*IC+JE-8 = IJ
        SACL    IJ

        ADLK    SGTBL           ;Points to slip gain table
        TBLR    DKA             ;Read DKA from table
        ADDK    1               ;Change pointer to read DKB
        TBLR    DKB             ;ACC = IJ+1
```

```
        ADDK    6               ;Change pointer to read DKC
        TBLR    DKC             ;ACC = IJ+7
        ADDK    1               ;Change pointer to read DKD
        TBLR    DKD             ;ACC = IJ+8

; ---------------------------------------------------------------------------
; ---------- DEFUZZIFICATION BY HEIGHT METHOD
; ---------- Get sum of truth value of fired rules (MRA+MRB+MRC+MRD)
        LAC     MRA
        ADD     MRB
        ADD     MRC
        ADD     MRD
        SACL    SUMMR           ;SUMMR = MRA+MRB+MRC+MRD

;---------- Get sum of products: MRi*DKi, i=A,B,C,D
        LT      MRA
        MPY     DKA
        LTP     MRB             ;ACC = MRA*DKA
        MPY     DKB
        LTA     MRC
        MPY     DKC
        LTA     MRD
        MPY     DKD
        APAC
        SACH    SUMPROD         ;SUMPROD=MRA*DKA+....+MRD*DKD

; ---------- Divide SUMPROD BY SUMMR do get DKs (Fractional division)
;       Here, the sign of the result is defined by SUMPROD only
;       since SUMMR  is by def. always positive.
        ZALH    SUMPROD
        ABS                     ;Make numerator positive
        RPTK    14
        SUBC    SUMMR           ;Result is in low ACC
        SACL    DKs
;
        LAC     SUMPROD
        BGEZ    DDONE2          ;Done if sign is positive
        ZAC
        SUB     DKs
        SACL    DKs             ;Negate quotient if negative

; ---------------------------------------------------
; ---------- GET NEW VALUE OF Ks (SLIP GAIN)
;
DDONE2:
        LDPK    6
        LAC     SWT2            ;Introduce a delay after FSG is set,
        BZ      PROC
        SUBK    1               ; to store rated condition into buffer.
        SACL    SWT2            ;Decrement counter SWT2
        BGZ     QUIT

PROC:   LDPK    7
        LT      GKs
        MPY     DKs
        PAC
        SACH    TEMPi
        ZALH    KsH             ;Use double precision to compute Ks.
        ADDS    KsL
        ADD     TEMPi           ;Add to low ACC since GKs = 0.64 bit.
```

```
; ---------- Impose min and maximum values for Ks, for safety reasons.
          ADDH    Ksmax           ;Ksmax = 32767 - maxKs (maxKs = 2Kso = 520)
          SUBH    Ksmax
          SUBH    Ksmin           ;Use saturation features to enforce min Ks.
          SUBH    Ksmin           ;Actually Ksmin = (32768+minKs)/2
          ADDH    Ksmin           ;Currently, minKs = 0.5 Kso = 130.
          ADDH    Ksmin

          SACH    KsH
          SACL    KsL
          LDPK    6
          SACH    Kslip           ;Kslip is the value used to get Wsl
QUIT:     RET
;
; ---------------------------------------------------
; ---------- DEFINE THE TABLE FOR DM's FOR Delta Ks
SGTBL: DW       dk2,    dk3,    0,      0,      dk5,    dk7,    dk8
       DW       dk1,    dk2,    0,      0,      dk5,    dk7,    dk8
       DW       dk1,    dk2,    dk4,    0,      dk5,    dk8,    dk7
       DW       dk1,    dk1,    dk2,    0,      dk5,    dk8,    dk8
```

VITAE

Gilberto Costa Drumond Sousa was born in Faria Lemos, MG, Brazil on October 5, 1958. He attended city elementary and middle schools and went to the Federal Technical School of Espirito Santo, Vitoria, ES, finishing his High School studies in 1976. He received a B.S. degree in Electrical Engineering from the Federal University of Espirito Santo, (UFES) in 1981. He joined the faculty of UFES as an Auxiliary Professor in 1982. Then he went to the Federal University of Santa Catarina for his Master studies and received a M.Sc degree in Electrical Engineering in 1986. He returned to UFES as an Assistant Professor and from 1986 to 1989 taught several undergraduate courses at the Electrical Engineering Department. In 1989 he received a four-year scholarship from the National Research Council and joined the University of Tennessee, Knoxville, for his doctoral studies. He received his Ph.D. degree in Electrical Engineering in December 1993.

The author has been involved in research areas including Power Systems and Industrial Controls, and more recently, Fuzzy Logic applications to Power Electronics and Drives. He has published several papers on the areas listed above.